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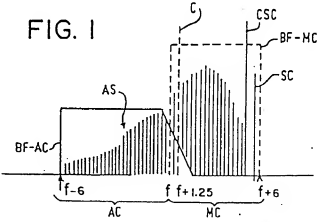
(54) System for broadcasting HDTV images over standard television frequency channels.

(57) Method and apparatus for transmitting and receiving an HDTV signal over standard television bandwidth channels. The system provides for quadrature modulation of the standard NTSC carrier with an augmentation signal containing components of the HDTV signal. The upper sideband portion of the quadrature modulation is suppressed along with the carrier produced from the quadrature modulation. The resulting lower sideband extends into the lower adjacent channel frequency spectrum, and is transmitted along with the standard NTSC video signal. Improved compatibility is achieved with existing NTSC signals by varying on an alternate line basis the phase of the augmentation signal to avoid the consequence of carrier phase shift from DC components in the quadrature modulated signal. A time gated demodulator is provided at the receiver for accurately tracking the phase of the carrier, permitting accurate demodulation of the quadrature augmentation signal. A demodulation circuit is described having phase shift compensation for removing the effects of phase delays incurred during processing of the received broadcast signal.

FIG. 1



EP 0 311 188 A2

SYSTEM FOR BROADCASTING HDTV IMAGES OVER STANDARD TELEVISION FREQUENCY CHANNELS

Field of Invention

The present invention relates to high definition television (HDTV) transmission systems. Specifically, apparatus and method are disclosed for broadcasting a standard NTSC television signal and augmentation video signal in adjacent standard television channels.

The United States Patent US-A 4,694,338 (PHA 21.323) describes techniques for increasing the resolution of images which are transmitted as video signals over standard broadcast channels. The enhanced image includes a component which can be used to increase the aspect ratio of the displayed image. Along with a transmitted main component, representing a standard NTSC video image, side panel information is transmitted which is suitably joined with the standard NTSC image to provide a wider aspect image for display. Additionally, provisions are made to transmit further horizontal and vertical detail for both the side panels and main image component in the augmentation channel.

The additional information transmitted in HDTV must be compatible with television receivers which are currently used to demodulate the standard NTSC broadcasts. Also, spectrum space must be efficiently used to preserve the interference protection presently afforded by guard bands between locally used channels so that information from one channel does not interfere with other broadcasts of different programs in other localities where these adjacent channel frequencies are used.

One proposed techniques for transmitting additional video image detail is set forth in United States Patent US-A 4,521,803 to Gittinger. This technique transmits an enhanced video image using quadrature modulation. Two horizontal lines of a high resolution picture are scanned and the luminance information of the two lines are added together to amplitude modulate a picture carrier in the standard NTSC format. An additional signal, however, is provided which is formed from the difference between the luminance signals of the scanned lines. This difference signal is used to form a suppressed, double sideband signal which is in phase quadrature with the standard NTSC carrier. The system transmits only the upper sidebands of both the in phase and quadrature phase modulations, limiting the bandwidth to the standard video bandwidth of an NTSC signal. In a second embodiment described in the patent, the in-phase modulation component is permitted to extend into the lower adjacent channel to form an increased bandwidth signal. Synchronous detection at the receiver is proposed to recover both the in-phase and quadrature phase modulated information. The system is proposed as being compatible with television receivers which are not equipped with special demodulation and decoding circuits.

In another quadrature system, the Matsushita Industrial Company has proposed during ICCE 1987 a scheme for quadrature modulation of an NTSC picture carrier. In this system, the lower sideband of both the quadrature and in-phase modulation components are filtered to limit the bandwidth to that of a standard broadcast channel. This system as well is believed to be compatible with existing television receivers, wherein it is assumed that the augmentation video signal, appearing as a quadrature component to the standard NTSC signal, will be transparent to currently used receivers which utilize quasi-synchronous demodulation techniques for recovering the baseband video signal.

Summary of the invention

It is an object of this invention to provide an improved, quadrature-based HDTV transmission system.

It is a more specific object of this invention to provide a quadrature-based system which has improved compatibility with existing television receivers.

In a first aspect of the present invention, use is made of the lower adjacent channels except in the cases of channels 2, 5, 7, 14 and 38, which are dedicated to non-broadcast services.

Use of this lower adjacent spectrum space is made possible by quadrature modulating the standard carrier for these channels with an augmentation signal which can be used in an appropriately equipped receiver to recreate the HDTV image. The augmentation video signal modulates a carrier signal having the same frequency as the standard NTSC broadcast, but in phase quadrature thereto. The resulting signal is a suppressed carrier double sideband modulated signal. The upper sideband of the quadrature phase carrier double sideband suppressed carrier signal is attenuated, using a filter at the transmitter which attenuates 6 db at the carrier frequency. In the preferred embodiment, this filter has an Nyquist (linear) slope function which slopes from the band edge of the primary channel negatively, and has a negligible loss to the lower sidebands of greater than 1.25 MHz. The filtered quadrature sidebands are linearly combined with

the standard NTSC carrier modulated signal.

The lower sideband of the augmentation signal modulated carrier is therefore transmitted in phase quadrature with the standard main channel NTSC signal. Thus, additional bandwidth is provided using the lower adjacent channel and a portion of the assigned channel. The possibility for cross-talk between the main and augmentation channels is lessened by suppressing the upper sideband of the augmentation modulated signal. The net result is an effective bandwidth of substantially 11 MHz for the HDTV transmission.

In a second aspect of the invention for improving the compatibility of HDTV broadcasts with standard NTSC video broadcasts, the effect of any low frequency component at or near DC in the augmentation signal on the carrier phase is lessened. A DC component contained in the quadrature modulated signal which is combined with the standard NTSC modulated signal, will tend to shift the carrier phase which is used at each of the conventional receivers for demodulation. Conventional receivers establish a phase reference with respect to the transmitted carrier signal. Any carrier shift resulting from a DC component contained in the augmentation signal will create an error in demodulating the modulated carrier signal by standard NTSC receivers having quasi-synchronous detectors of the usual kind. The net effect is cross-talk which would enter these standard television receivers due to the presence of the quadrature modulated signal. To remove the effects of this DC component, it is proposed that the phase of the carrier carrying the augmentation signal be reversed 180° on alternate line during transmission of those signal components having a DC component. This will effectively remove the effect of a DC component on the recovered in-phase picture carrier.

In still a further improvement according to the invention, accurate carrier regeneration is provided at the receiver equipped with circuitry for demodulating the quadrature components contained in the HDTV signal. Accurate carrier regeneration is provided by a technique which samples the phase of the carrier during specific time periods, corresponding to a portion of the blanking interval of the video signal. During at least a portion of the horizontal blanking intervals of the standard NTSC video signal, the quadrature components at the transmitter are fully suppressed, providing for an accurate representation of the transmitted carrier phase and frequency. Thus, carrier lock is achieved only during the blanking time when the true phase of the carrier is known. The remaining portion of the horizontal blanking interval may be used for data transmission.

In still a further improvement provided by the invention, the demodulating circuit for detecting the quadrature modulated augmentation signal includes a technique for removing the phase offset incurred during processing of the signal. The present invention provides for a demodulation circuit which will alter the phase of the locally generated carrier to compensate for any phase delays incurred by the intermediate frequency signal during signal processing. Thus, the intermediate frequency signal is accurately demodulated with respect to its phase to accurately recover the baseband augmentation signal.

According to a first broad aspect of the invention, there is provided a method for transmitting an augmentation video signal for increasing the width and information content of a standard picture represented by a standard video signal comprising:
generating a radio frequency carrier signal for a channel of the standard television broadcast spectrum;
modulating said carrier signal with said standard video signal to produce an amplitude modulated carrier signal;
modulating said carrier signal with said augmentation video signal to produce an amplitude modulation component in quadrature with modulation produced by said standard video signal;
suppressing the upper sideband of said quadrature amplitude component, whereby substantially all of said quadrature amplitude components are within a bandwidth allotted to a lower adjacent channel; and,
switching the phase of said quadrature components 180° during alternate lines of said standard video signal, whereby the effects of a DC component in said augmentation signal on the mean phase of said carrier phase is removed.

According to a second broad aspect of the invention, there is provided a method for demodulating an augmentation signal and standard video signal modulated on a carrier signal of a standard broadcast channel, comprising:
generating an intermediate frequency signal from said modulated carrier signal;
generating a local carrier for demodulating said broadcast channel by establishing a phase of a voltage controlled oscillator during a blanking interval of said standard video signal by applying a control voltage proportional to the difference in phase between said voltage controlled oscillator signal and the phase of said intermediate frequency signal, whereby said oscillator assumes a constant phase with respect to said intermediate frequency signal;
compensating said local carrier for phase delays incurred by said intermediate frequency signal comprising:
determining during a horizontal blanking interval of

said video signal the difference between the phase of said local carrier and a reference phase voltage;
shifting the phase of said local carrier signal in a direction to reduce said difference; and,
phase demodulating said intermediate frequency signal with said shifted local carrier signal, whereby said augmentation signal is produced.

According to a third broad aspect of the invention, there is provided a method for transmitting an augmentation video signal for increasing the width and resolution of a standard video signal comprising:
generating in phase and quadrature phase components of a carrier signal of a television channel;
modulating said in-phase component of said carrier signal with a standard NTSC video signal;
modulating said quadrature component of said carrier signal with said augmentation video signal, whereby upper and lower sideband components of said quadrature component are produced; and
combining said lower sideband of said quadrature component containing said augmentation video signal with said in-phase modulation components, whereby said modulated carrier signal is produced having a quadrature component in a lower adjacent channel containing said augmentation signal.

According to a fourth broad aspect of the inention, there is provided an apparatus for transmitting an augmentation video signal along with a standard video signal comprising:
a carrier generator for generating a carrier signal having in-phase and quadrature phase signal components at a frequency of a standard television channel;
modulator means for modulating said in-phase component with said standard video signal and said quadrature phase component with said augmentation video signal;
means for suppressing an upper sideband produced by modulating said quadrature component, whereby only a lower sideband produced by modulating said quadrature component remains: and,
means operative during a blanking interval of said video signal for suppressing all quadrature related modulation components, whereby only in-phase signal components are produced during said blanking interval.

According to a fifth broad aspect of the invention, there is provided in a system for producing high definition television signals which include a standard video signal transmitted as an upper sideband, and an augmentation video signal, transmitted as a lower sideband of a single carrier frequency, a receiving apparatus for recovering said augmentation video signal comprising:
means for converting a received carrier frequency signal containing an upper and lower sideband into an intermediate frequency signal which contains an unmodulated component and in-phase and quadrature phase components;
means for generating a local carrier signal having a frequency and phase locked to the frequency and phase of said unmodulated component contained in said intermediate frequency signal;
means for phase shifting said local carrier signal to obtain a substantially 90° phase relationship with quadrature modulation components contained in said intermediate frequency signal; and
means for phase demodulating said quadrature components using said local carrier.

According to a sixth broad aspect of the invention, there is provided a method for transmitting an augmentation video signal for increasing the width and resolution of a standard video signal comprising:
generating in-phase and quadrature phase components of a carrier signal of a television channel;
modulating said in-phase component of said carrier signal with a standard NTSC video signal;
modulating said quadrature component of said carrier signal with said augmentation video signal, whereby upper and lower sideband components of said quadrature component are produced;
combining at least one of said quadrature components containing said augmentation video signal with said in-phase modulation components, whereby said modulated carrier signal is produced having at least one quadrature component containing said augmentation signal; and,
alternating the phase of said combined quadrature components 180° on an alternate periodic basis, whereby the effects of a DC component in said augmentation signal on said carrier signal phase are reduced.

Brief description of the Figures

Figure 1 illustrates the frequency spectrum of an HDTV transmission signal in accordance with a preferred embodiment of the invention.

Figures 2A illustrates the amplitude relationship between the carrier, upper and lower sidebands in an IF signal of an NTSC receiver.

Figure 2B illustrates the vectoral relationship between the upper an lower sidebands of the NTSC transmitter signal.

Figure 2C illustrates the effect due to the upper and lower sidebands on an NTSC signal.

Figure 2D illustrates the effect of the quadrature transmitter signal on the phase of the NTSC transmitted carrier, as well as the effect of line sequential alternation of the quadrature signal.

Figure 3A illustrates the intermediate frequency passband frequency response of a standard NTSC television receiver.

Figure 3B illustrates the frequency response of the passband filter for the quadrature components of the quadrature generated signal of Figure 1.

Figure 4A illustrates a standard NTSC video line signal.

Figure 4b illustrates the augmentation signal amplitude versus time with respect to the standard NTSC video signal.

Figure 5 illustrates a transmitting apparatus for transmitting the signal whose frequency spectrum is shown in Figure 1.

Figure 6 demonstrates a receiver carrier regeneration circuit and NTSC demodulator for recovering the NTSC signal whose spectrum is shown in Figure 1.

Figure 7 illustrates a carrier demodulator for the quadrature component of the signal whose frequency spectrum is shown in Figure 1.

Figure 8 is a timing diagram illustrating the timing pulses derived from the NTSC signal for enabling the gate elements of Figures 6 and 7.

Description of the Preferred Embodiment

Referring now to Figure 1, there is shown an amplitude versus frequency spectrum plot of a broadcast signal produced in accordance with the present invention. The spectrum includes a standard NTSC modulated signal in an upper channel, referred to in the following as the main channel MC. The carrier C is located at 1.25 MHz above the lower band edge for this channel. The main channel MC occupies a bandwidth of 6 MHz, as shown by its bandwidth function BF-MC. The standard color subcarrier CSC and sound carrier SC are shown within the main channel containing the standard NTSC broadcast spectrum.

The additional information transmitted in HDTV must not interfere with the operation of NTSC receivers presently serving the public. As the radio frequency spectrum is a valuable public resource, the amount of additional spectrum required for HDTV should be held to a minimum. One possibility of meeting these requirements is to transmit the additional information in the lower adjacent TV channel AC as those frequencies are not used for broadcasting in the same or nearby communities, and presently serve in a guardband locally. Use of these foquencies as other programming channels, if possible, would increase the efficiency of the spectrum utilization, however, new and novel modulation techniques would be required to permit continued operation of NTSC receivers in those communities.

Additional to the standard NTSC broadcast spectrum, there is shown an augmentation signal spectrum AS which comprises a frequency spectrum extending into the lower adjacent channel AC. The amplitude of the spectrum, indicated by its bandwith function BF-AC, is generally limited to be 10 dB below the broadcast carrier level. The additional augmentation signal sideband is in phase quadrature to the carrier, and to the main channel sideband information.

Modulating a carrier signal having the same frequency as the standard broadcast signal but in phase quadrature thereto with the augmentation signal produces a set of sidebands symmetrical with the broadcast carrier. These augmentation signal sidebands and their suppressed quadrature carrier are filtered so that most of the upper sideband energy is suppressed at 1.25 MHz above the picture carrier, the picture carrier is attenuated 6 dB and the spectrum 1.25 MHz below the carrier is not attenuated.

The filter function from the upper band edge of the lower adjacent channel and main channel decreases linearly with frequency. The picture carrier frequency is attenuated 6 dB. The filter slope remains linear up to the stop band which is 1.25 MHz above the carrier frequency. A second stop band exists at the lower bandwidth edge of the lower adjacent channel. The frequency spectrum produced from the augmentation signal is therefore confined to its lower sideband having most of its energy below the carrier frequency. Only a small portion of the upper sideband energy remains in the frequency spectrum between the carrier frequency and 1.25 MHz above the carrier frequency.

The Nyquist filter amplitude response is generally shown in Figure 3B. Figure 3B illustrates a Nyquist response which has a negatively decreasing function which is linear, beginning at the band edge between the upper and lower channels. The amplitude further at the carrier frequency is 6 dB attenuated from the band edge amplitude level. The standard NTSC television receiver generally has an IF signal passband as is shown in Figure 3A. The frequency amplitude function in the region of the carrier or corresponding intermediate frequency signal, is shown to be generally complementary to that of Figure 3B. As can be seen in Figure 3A, the lower sidebands of the intermediate frequency signals are subject to an attenuation which is greater than those of the upper sideband. The IF signal passband of Figure 3A will include a quadrature component representing the augmentation signal having an upper and lower sideband which are substantially equal. It is necessary to filter the quadrature RF signal according to Figure

3B before transmitting the same in order to produce symmetrical quadrature sidebands for the received quadrature signal after filtering by the intermediate frequency signal passband filter. Quasi-synchronous and true synchronous detectors will reject the present quadrature signal sidebands which appear with the in-phase NTSC IF signal sidebands.

Television receivers employed envelope detection of the intermediate frequency picture signal before the development of quasi-synchronous video detector integrated circuits some twelve years ago.

The quasi-synchronous videodetector differs from an envelope detector in that it synchronously detects the modulation by regenerating the carrier frequency (at IF) and it is this regenerated carrier which controls the conduction of the detecting means. The bandwidth of this carrier regeneration circuit is typically a few hundred kilohertz, although it could be made arbitrarily smaller. Quasi-synchronous video detectors are now in use in North America. True sychronous video detectors have much smaller bandwidth in their carrier recovery circuit. It is expected that any present usage of envelope detection will soon end due to the economic advantages of the newer detection process.

Compatibility of the proposed scheme with NTCS receivers is therefore compatibility with receivers having quasi-synchronous video detectors, and not those few older receivers still in service having envelope detectors. It is thought that such receivers will be out of service by the time a new system of broadcasting can be implemented.

The carrier regeneration process in quasi-synchronous detectors must provide a carrier signal to the detection means at the correct phase to demodulate the wanted information. This means in the case of NTSC receivers, the carrier must be in-phase with the NTSC component, and would therefore be in quadrature with the additional signal transmitted on the same carrier frequency for HDTV.

The augmentation information, which is in phase quadrature with the standard NTSC video signal, is compatible with existing television receiving sets having quasi-synchronous video demodulator circuits. No additional carrier component is transmitted. The augmentation signal is produced by a balanced modulator which fully suppresses the carrier. Most of the upper sideband is suppressed as well. The resulting lower sideband is combined linearly with the standard NTSC broadcast signal to provide the spectrum of Figure 1. Standard television receivers equipped with quasi-synchronous demodulation detectors will not detect the additional quadrature component containing the augmentation signal. Only a very low level of power is introduced into the lower adjacent channel spectrum by the higher order lower sidebands, and will not constitute interference with other stations in adjacent localities. Power in the sidebands contained from augmentation signal modulation of the carrier decreases rapidly as the frequency separation from the carrier increases, thus reducing the potential interference into a signal broadcast in the lower adjacent channel.

The picture signal according to NTSC standards modulates the carrier approximately 58% depth of modulation, white being 12 1/2% and black 75%.

The quadrature signal being transmitted uses balanced modulation with zero amplitude at 50 IRE scale units (mid-gray) and 29% carrier representing both black and white (but with opposite phase). Furthermore, the highest amplitude sidebands produced by the modulation process are attenuated at the transmitter by 6 dB. Taking these factors into account, it is seen that the power radiated in the lower adjacent channel (for equal signal-to-noise ratios) is small and is generally a function of the modulating video signal. It should also be noted that there is no aural carrier radiated in the lower channel.

Further reduction in possible interference with reception of the broadcasts on the lower adjacent channel in distant cities can be obtained by means of improved antenna directivety, where the HDTV signal is broadcast with circular polarization. In such cases, a circularly polarized receiving antenna can further attenuate the unwanted lower sideband of the HDTV transmission.

Referring to Figure 2A, the frequencies and amplitudes of both sidebands of a vestigial sideband signal at the output of the IF filter in the receiver is shown. This represents the case for baseband frequencies below 1.25 MHz which amplitude modulate the picture carrier in a standard NTSC signal. The frequency above the carrier in the upper sideband is denoted USB, and the frequency below the carrier in the lower sideband is denoted LSB.

The phase relationship of these frequencies USB and LSB with respect to the picture carrier phase CP defined by the horizontal axis is shown in Figure 2B. The carrier phase CP remains stationary, while the phase vector USB rotates counterclockwise and the phase vector LSB rotates clockwise at an angular velocity determined by the frequency separation between the sideband components and the carrier frequency. The sum of these components is shown by the vector SUM which is a phase vector rotating at the same angular frequency, but whose amplitude traces out the elliptical path shown.

For a standard NTSC signal, the vector SUM of

Figure 2B is added to the carrier phase vector CPV as shown in Figure 2C. The carrier amplitude and phase as a function of time lies on the elliptical path shown. The carrier CDCM for DC modulation is shown by the dashed vector which lies on the carrier phase axis with no phase offset at any time. Note that for frequencies above DC, the amplitude modulation CAM is large (major axis of ellipse) while the phase modulation is small (minor axis of ellipse). Thus, an essentially constant phase amplitude modulated signal is produced.

The quadrature phase, suppressed carrier signal produces a similar pair of inequal amplitude sidebands which is the mirror image of those shown in Figures 2A and 2B, with the carrier phase shifted 90° with respect to the NTSC main channel picture carrier. Thus, Figure 2B is rotated 90° to represent the locus of all phases of the quadrature modulation signal with respect to the main channel carrier. The resultant carrier amplitude and phase for the picture carrier with quadrature vestigial sideband modulation is shown in Figure 2D. The carrier for DC modulation in the quadrature channel is shown by the dashed vectors which lie perpendicular to the carrier phase axis. The upper dashed vector V1 represents the DC induced phase component of the quadrature signal. The bottom dashed vector V2 illustrates the effect of shifting the quadrature added components 180°. The net result is an average phase which has not changed. Thus, a constant phase error of the resultant main channel picture carrier is produced.

The constant phase of the resultant main channel picture carrier and a quadrature modulation signal with a 180° phase inversion of the quadrature suppressed carrier is also shown in Figure 2D. The same phase error magnitude results, but in the opposite angular direction from the main channel in-phase picture carrier.

In order to avoid the shift in carrier phase which would accompany a DC component contained in the quadrature component, the present invention proposes to switch the carrier phase quadrature added component 180° on alternate lines of the video signal. Thus, the average carrier phase error due to any DC component would be effectively average to zero. As is shown in Figure 2D, switching the lower sideband quadrature component 180° tends to pull the carrier phase in an opposite sense during alternate lines of video augmentation signal. The average phase for the carrier, when switching the quadrature added component 180° is therefore substantially zero. By providing for alternate line phase inversion of the quadrature added component, the effects of DC components contained in the augmentation signal sideband on carrier phase are minimized. Standard television receivers equipped with quasi-synchronous detectors will therefore see an accurate carrier phase with which to generate the necessary reference signal for demodulating the standard NTSC upper sideband components.

Turning now to Figures 4A and 4B, there is shown one line of a standard NTSC video signal amplitude versus time function, and a line of augmentation video signals. The familiar blanking interval BI containing a sync pulse and colorburst are shown, followed by an active line portion ALP of 52 microsseconds. During the active line portion ALP, the luminance information for display on the CRT screen is used to modulate the beam intensity of a cathode ray tube. Additionally, an augmentation signal is transmitted and received at the same time, containing left panel LP, right panel RP, LD/HD and digital audio DA information. The left and right panels are, of course, the remaining video information comprising the additional width for a wide aspect picture. The additional width information is contained in 18 microseconds per line interval and can be joined to a standard NTSC video frame to provide a wider aspect ratio picture. Additionally, vertical and horizontal resolution components are transmitted as LD and HD components, followed by a digital audio signal DA which can be used with the standard NTSC audio signal to provide improved sound quality for television, including stereophony.

The present invention may provide for a suppression of all quadrature components contained in the spectrum of Figures 1 and 2 during at least a portion S of the blanking interval of each NTSC video signal. It is also possible to transmit information I during the remaining portion of horizontal blanking intervals using the quadrature channel, provided that the information is bandpass limited between 1.25 and 7.25 MHz and its mean value is 50 IRE, at which level the modulator 33 output is zero. Thus, except for such data signals, during blanking the only transmitted signal is an NTSC standard video signal having an in-phase carrier and in-phase sideband components. This interval is advantageously used in a preferred embodiment of the invention to generate an accurate carrier regeneration at the receiver since the true phase of the carrier is known when quadrature components to which the NTSC channel may be responsive, that is, components below 1.25 MHz at baseband are fully suppressed. The lower sideband of the quadrature signal may convey information at frequencies above the effective bandwidth of the carrier recovery circuitry during the horizontal blanking period without any adverse effect upon the accuracy of the carrier recovered. Additionally, during the left panel and right panel time, and during transmission of any component having a DC component, the effect of the DC component is avoided

by switching the phase of the quadrature components on an alternate line basis 180° for the quadrature modulated augmentation signal.

Having generally described the nature of the signal which is produced in accordance with the transmission system of the present invention, reference may be made to Figure 5 which illustrates a technique for generating the signals of Figures 1 and 2. An HDTV source 10 is shown, connected to a transcoder 11. The HDTV source 10 may produce an 1125 line, 60 field per second, 2 to 1 interlaced studio standard television signal. Transcoder 11 would convert the HDTV source signal 10 to that required by encoder 12. The encoder 12 will receive R, G, B inputs of 525 lines per picture, progressive scan. The non- interlaced video information provides a picture at 59.94 fields per second, with an aspect ratio of 16 to 9.

The encoder 12 also provides horizontal blanking pulses BP identifying the horizontal blanking time for each line of video signal being transmitted. The horizontal blanking pulses occur at the 525 line rate, synchronized with the NTSC signal produced by encoder 12.

One of the outputs of the encoder 12 is a standard NTSC video signal SV of Figure 4A, which is applied to a modulator 15. A sideband filter 20 suppresses the lower sideband of the NTSC modulated carrier signal supplied by carrier generator 14. The resulting signal from sideband filter 20 lies within the allocated bandwidth of a main NTSC channel containing standard NTSC video program information.

The augmentation signal AS supplied from the encoder 12 which may be in the format of Figure 4B is AC coupled through a capacitor 21. Gate 22 will DC-restore the video signal during a portion of the blanking interval. During this portion of the blanking interval, established by monostable multivibrators 26 and 27, or other suitable timing means, the capacitor 21 is briefly connected to ground level, providing DC restoration of the augmentation video signal AS. This DC restoration time occurs for a pulse width of approximately 3 microseconds, as established by the monostable multivibrator 27. Multivibrator 27 is triggered after a delay of 1 microsecond generated by monostable multivibrator 26. Monostable multivibrator 26 is in turn triggered by the leading edge of the horizontal blanking interval which is synchronized with signals from the encoder 12. The resulting clamping action provides for a DC voltage level of zero volts to the input of transmission gate 28. This corresponds to the black video level, 0 IRE units in the standard IRE scale, where white is 100 IRE and mid-gray is 50 IRE. During the active line time, which occurs between horizontal blanking intervals, the active video signal is fed through transmission gate 28 to the input of balanced modulator 33. During the blanking time, however, transmission gate 28 is non-conductive by virtue of its control inputs being connected to the horizontal blanking pulse.

Additionally, during the horizontal blanking time for the video signal, transmission gate 32 is operative so that modulator 33 is fed with a reference level, established by potentiometer 34 to establish a 50 IRE video signal level for application to the balanced modulator input 33. This constitutes a bias level for modulator 33 so that during blanking, the modulator output is fully suppressed. During the active line portion, between blanking intervals, video levels above 50 IRE produce an output in phase with the input carrier signal, whereas video levels below 50 IRE produce an output phase shifted 180° from the input carrier signal phase.

If it is desired to transmit data during the horizontal blanking interval, said data signal may be coupled by capacitor 36 and resistance 35 to gate 32 from the data source 37. Said data spectrum must lie in the frequency range above 1.25 MHz and below 7.25 MHz.

Digital data must be channel coded so that below 1.25 MHz there is no significant spectral component. Such a component, if present, would cause a phase error in quasi-synchronous NTSC receivers. The data must be confined in a fixed portion of the horizontal blanking interval. The remaining portion of the horizontal blanking interval is left free of any quadrature signal to permit a phase calibration interval for the augmentation signal demodulated at the users receiver. This data transmission portion of the blanking interval is achieved by synchronizing the data stream produced by data source 37. The data source 37 may be configured to produce a data stream in the later portion of the horizontal blanking inteval, leaving the earlier portion free for calibration purposes. Biphase of Manchester coding of the data would provide the necessary spectral shaping.

The input carrier signal phase is controlled by a doubly balanced modulator 16 connected through an input terminal to a phase shifter 13. Phase shifter 13 will provide a 90° phase shifted carrier, constituting a quadrature carrier having the same frequency as the standard NTSC signal produced by modulator 15. This quadrature related carrier signal is applied through modulator 16 and gate 18 as an RF input signal to modulator 33.

The phase of the output of doubly balanced modulator 16 is controlled by a switching signal applied from a divider 24 and NAND gate 30 to an inverter 25. During the active portion of even numbered video lines, the output signal of modulator 16 is in phase with its input. This also results during horizontal blanking intervals. During the active portion of odd numbered video lines, the output of

doubly balanced modulator 16 is 180° out of phase with its input. The alternate line phase shifted carrier signal is received by transmission gate 18. Means 18 may be operationally convenient to assist in checking for carriersuppression by disabling said means manually through switch 17 and noting whether there is any change in carrier power during blanking. The quadrature modulated carrier signal produces by modulator 33 is applied to a Nyquist sideband filter 28. Sideband filter 38 is selected to suppress the upper sideband set produced by modulator 33, leaving the lower sideband set which resides in the lower adjacent channel. This filter 38 attenuates the upper sideband frequencies produced by the augmentation signal AS. The filter has an amplitude response, as is shown in Figure 3B. The filter response is selected so that virtually all upper sideband components which lie 1.25 MHz above the carrier frequency are eliminated. The Nyquist filter function permits quadrature-produced modulation components below the band edge frequency to pass with negligible attenuation. The filter characteristic for the Nyquist filter is selected to have a linear decreasing slope, beginning at the band edge between upper and lower adjacent television channels. The filter response at the picture carrier frequency is 6 dB down from the band edge. The other stop band for filter 38 is at the lower band edge of the lower adjacent channel.

Tho filtered quadrature sideband signal is combined in signal combiner 23 to produce the composite RF video signal spectrum as is shown in Figure 1. This, of course, may be applied to a frequency converter to up or down convert the signal to the desired broadcast channel frequency.

Having thus described the nature of the signal, and a technique for generating the signal for transmitting the HDTV signal in accordance with the invention, a receiving apparatus suitable for demodulating the transmitted signal will be described with respect to Figure 6.

Referring now to Figure 6, there is shown a receiving circuit for demodulating the NTSC video signal. The circuit of Figure 6 includes a carrier regenoration circuit which providos a local carrier signal at the nominal picture IF frequency, which is in phase with the transmitted NTSC carrier signal. The output of the carrier regeneration circuit is applied to an augmentation signal demodulator ASD shown in Figure 7. The carrier regeneration circuit of Figure 6 is a phase locked loop having a voltage controlled oscillator 76. The loop is responsive to an error signal from DC to a cut-off frequency determined by low pass filter 71 or 72. In order to avoid the consequences of a significant DC component as a result of the quadrature modulating signal, a time gated froquency control is implemented.

The intermediate frequency signal produced by a television receiver equipped to receive HDTV broadcasts is applied to a first intermediate frequency amplifying stage 45. This amplified intermediate frequency signal is applied to gate 46. Gate 46 is conductive during the latter portion of the horizontal blanking interval, as well as during the active line time. The gated intermediate frequency signal is applied to the Nyquist filter 48 having the general frequency characteristic shown in Figure 3A, and to the augmentation signal demodulator ASD of Figure 7.

The gated intermediate frequency signal is known to have a true phase during the horizontal blanking interval with respect to the transmitted NTSC carrier and upper sideband. Product detectors 56 and 57 will demodulate the intermediate frequency signal using the local carrier generated from voltage controlled oscillator VCO 76. The phase shift 90° is introduced between the re ferecce inputs of each product detector 56 and 57 to provide an in-phase component corresponding to the NTSC baseband signal BS, and a quadrature phase component which provides the error signal for the voltage controlled oscillator 76.

The control voltage for the voltage controlled oscillator 76 is applied through two gates 73 and 74. Gate 74 is operative when the voltage controlled oscillator 76 is determined to be in carrier phase synchronization with the transmitted NTSC signal components. Gate 73 is operative during acquisition of the phase synchronization between voltage controlled oscillator 76 and the NTSC carrier signal. During a carrier acquisition stage, the filter 72, having a wider bandwidth than that of 71, will permit the phase locked loop sufficient bandwidth to locate and lock to the transmitted NTSC picture carrier.

Control over the bandwidth or the phase locked loop structure of Figure 6 is implemented by sampling the NTSC baseband signal and integrating the samples. As the level of this signal increases to a threshold for flip flop 70, the flip flop will be operative to switch transmission gates 73 to an OFF condition, and 74 to an ON condition, wherein the narrower filter 71 feeds a control voltage to voltage controlled oscillator 76. Sampling of the NTSC baseband demodulated signal is accomplished as will be evident with respect to Figure 8, only during the interval corresponding to the second portion of the blanking interval for the standard NTSC video signal. As was explained with respect to Figures 3A and 3B, the NTSC carrier transmitted during the horizontal blanking interval is free of any quadrature components greater than 1.25 MHz. Additionally, the NTSC carrier signal is at its maximum amplitude during the horizontal blanking inter-

val while any noise is constant in amplitude. Thus, it is possible for the carrier regeneration circuit of Figure 6 to lock and establish the true carrier phase at this time with ample signal amplitude to correctly synchronize. During the remaining portion of the video signal, gates 60 and 64 are disabled so that the VCO 76 remains locked at a carrier phase determined during the latter portion of each horizontal blanking interval.

The error voltages obtained during this sampling time are stored on capacitors 66 and 68. Amplifiers 67 and 69 provide a buffer function for these stored voltages.

Video signals coupled through capacitors 61 and 62 are DC restored by gates 59 and 63 during the first portion of the blanking interval.

Referring to Figure 8, the various signals necessary to provide DC restoration and sampling are shown. These signals are derived from the NTSC video signal (waveform a) produced from the NTSC sychronous video detector 56. A sync separator 51 will provide the horizontal sync show in waveform b of Figure 8, as well as a vertical synchronization pulse. Deflection circuitry 53, in the conventional manner, will produce horizontal deflection pulses (waveform c) at twice the horizontal sync pulse interval rate and synchronized thereto. As the HDTV system operates on a scan rate twice the conventional NTSC scan rate, horizontal deflection pulses necessarily are at this enhanced scanning rate.

A pulse generator 52 operates from the horizontal deflection pulses 54. Pulse generator 52 will divide the horizontal deflection rate (54) by two as shown in waveform d of Figure 8. The divided horizontal deflection pulse rate is used to generate a waveform e which represents the period of the NTSC horizontal blanking interval which occurs between the leading sync pulse edge, and end of blanking interval. These pulses have been designated PO. Additionally, a gate time P3 is generated which occurs between the leading edge of the horizontal synchronization pulse and the end of horizontal blanking time. As can be seen from waveform f, this occurs at an alternate rate, corresponding to the even fields of the video signal. Thus, it is possible to accurately mark that portion of the video signal having the alternate line phase reversal using P3, as described with respect to Figure 4B. ϕ indicates non-inverted phase and ϕ indicates inverted phase.

Waveforms g and h of Figure 8 illustrate pulses P1 and P2, identifying that portion of the blanking interval measured between the leading edge of the sync pulse and midway through the blanking interval. P2 illustrates a pulse which begins at the trailing edge of P1 and ending when the horizontal blanking poriod ends.

These pulses, as applied to those gates shown in Figure 6, permit the IF signal to pass through gate 46 during the NTSC video signal time, excluding that occupied by pulse P1. Additionally, gates 59 and 63 provide DC restoration during P1. The remaining portion of the horizontal blanking interval, as identified by pulse P2 will provide for a sampling of the NTSC in-phase and quadrature phase components via product detectors 56 and 57. The quadrature component is used to form a phase control voltage for VCO 76, while the in-phase component (the NTSC video signal) is also integrated and used to control the bandwidth of the phase locked loop formed from voltage control oscillator 76 and product detector 57.

Having thus illustrated how a local carrier regeneration is accomplished in a receiving apparatus for HDTV television receivers, reference may be had to Figure 7 wherein a demodulator for the quadrature augmentation signal is shown. The quadrature augmentation signal is also demodulated using synchronous detection techniques. However, the quadrature augmentation signal component contained in the baseband IF signal will suffer phase delays due to signal processing in the IF stages of the receiving apparatus. It is therefore required to adjust the reference carrier used for demodulating the quadrature augmentation signal so that it represents the true phase of the original carrier signal transmitted by the apparatus of Figure 5.

The circuitry of Figure 7 first includes an IF filter 79 which has an upper stop band approximately 1.25 megacycles above the carrier. The lower stop band for the signal filter 79 is located at the lower edge of the lower adjacent channel. The filtered signal is applied to first and second product detectors 80 and 81. Each of these product detectors are fed with a reference signal to demodulate a component of the augmentation signal contained in the IF signal.

The reference carrier frequency use for this demodulation is obtained from the VCO 76 of Figure 6. The reference carrier is accurately adjusted in phase by using amplitude modulators 93 and 94. These modulators will function as an attenuator unit, combining a quadrature component from phase shifter 96 with an in-phase component of the reference carrier, RC. By adjusting the relative levels of the signals produced by modulators 93 and 94, it is possible to adjust the phase of the resulting reference carrier signal RC. A summing network 97 will combine the outputs of modulators 93 and 94 to provide the reference carrier having a phase selected in accordance with the relative signal magnitudes from each of the modulators 93 and 94. The resulting reference signal is supplied to a balanced modulator 99 which will be under

control of pulse P3 of Figure 8, thereby providing alternate-line-phase-reversal-of-the-carrier-reference signal to each product detector 80 and 81.

The adjustment of the phase for the reference carrier is carried out during the interval defined by pulse P2 of Figure 8. During the portion of the horizontal blanking interval, identified by P1 of Figure 8 occurring between the leading edge of the sync pulse and within 3 microseconds (indicated by L-3) thereafter, DC restoration is provided by gates 86 and 91 for video signals coupled through capacitors 83 and 84. The following portion of the blanking interval as identified by pulse P2 will sample the outputs from each of the product detectors 80 and 81. The sampled outputs are applied to integrating circuits 88 and 89. As will be recalled from discussions of Figures 3A and 3B, the phase of the received carrier signal during the second portion of the blanking interval identified by P2 is that of the NTSC carrier, as all quadrature components less than 1.25 MHz are suppressed. Thus, the true phase of the received carrier signal, after conversion to the IF signal and being filtered through the IF filter 79 is represented by the outputs of product detectors 81 and 80. The proper carrier phase for recovering the quadrature modulated signal is 90° from the NTSC signal. Using this relationship, the NTSC signal can be shifted 90° to generate the required reference signal.

The portion of the blanking interval, P2, contains no quadrature signal components below 1.25 MHz. Components above 1.25 MHz are outside the bandwidth required of the phase controlled circuit, and produce no detectable phase error. The voltages derived during this interval of time having been integrated by integrators 88 and 89 are applied to control the signal levels for balanced modulators 93 and 94 in a magnitude such as to reduce the total phase error produced by product detectors 80 and 81. P3 will control phase switching of modulator 99. Thus, accurate phase regeneration for product detectors 80 and 81 is achieved during that portion of the blanking interval during which the phase is accurately known. P3 will switch the phase of the reference signal to product detectors 80, 81 during the active line portion of odd lines providing reinversion of odd line periods.

Thus, there has been described a receiving apparatus for regenerating a local carrier signal, as well as for demodulating the quadrature augmentation signal contained in the HDTV broadcast. Those skilled in the art will recognize yet other variations and embodiments of these techniques for implementing the invention.

Claims

1. A method for transmitting an augmentation video signal for increasing the width and information content of a standard picture represented by a standard video signal comprising:
generating a radio frequency carrier signal for a channel of the standard television broadcast spectrum;
modulating said carrier signal with said standard video signal to produce an amplitude modulated carrier signal;
modulating said carrier signal with said augmentation video signal to produce an amplitude modulation component in quadrature with modulation produced by said standard video signal;
suppressing the upper sideband of said quadrature amplitude component, whereby substantially all of said quadrature amplitude components are within a bandwidth allotted to a lower adjacent channel; and,
switching the phase of said quadrature components 180° during alternate lines of said standard video signal, whereby the effects of a DC component in said augmentation signal on the mean phase of said carrier phase is removed.

2. The method of claim 1 wherein said carrier signal is phase reversed only during portions of said augmentation signal which contain a DC level.

3. The method of claim 1 further comprising filtering said augmentation signal so that the amplitude of said lower sideband containing said augmentation signal components decreases near said carrier frequency as a substantially linear function.

4. The method of claim 1 further comprising suppressing said quadrature component bearing said augmention video signal during said horizontal blanking period.

5. A method for demodulating an augmentation signal and standard video signal modulated on a carrier signal of a standard broadcast channel, comprising:
generating an intermediate frequency signal from said modulated carrier signal;
generating a local carrier for demodulating said broadcast channel by establishing a phase of a voltage controlled oscillator during a blanking interval of said standard video signal by applying a control voltage proportional to the difference in phase between said voltage controlled oscillator signal and the phase of said intermediate frequency signal, whereby said oscillator assumes a constant phase with respect to said intermediate frequency signal;
compensating said local carrier for phase delays incurred by said intermediate frequency signal comprising:
determining during a horizontal blanking interval of

said video signal the difference between the phase of said local carrier and a reference phase voltage; shifting the phase of said local carrier signal in a direction to reduce said difference; and. phase demodulating said intermediate frequency signal with said shifted local carrier signal, whereby said augmentation signal is produced.

6. The method of claim 5 wherein said reference phase voltage is the intermediate frequency signal during said horizontal blanking period.

7. The method of claim 5 further comprising alternately phase shifting said local carrier signal 180°.

8. The method of claim 5 further comprising holding the phase of said local carrier signal constant between blanking intervals.

9. A method for transmitting an augmentation video signal for increasing the width and resolution of a standard video signal comprising:
generating in phase and quadrature phase components of a carrier signal of a television channel;
modulating said in-phase component of said carrier signal with a standard NTSC video signal;
modulating said quadrature component of said carrier signal with said augmentation video signal, whereby upper and lower sideband components of said quadrature component are produced; and
combining said lower sideband of said quadrature component containing said augmentation video signal with said in-phase modulation components, whereby said modulated carrier signal is produced having a quadrature component in a lower adjacent channel containing said augmentation signal.

10. The method of claim 9 further comprising suppressing said quadrature carrier component as well as said upper sideband component produced by said augmentation signal.

11. The method of claim 9 further comprising suppressing changes in the mean phase of said quadrature signal produced by a DC component contained in said augmentation signal.

12. The method of claim 9 wherein said upper sideband is suppressed using a filter having a reverse Nyquist response.

13. The method of claim 12 wherein said filter has an amplitude versus frequency response which decreases linearly beginning at a frequency at an upper edge of said lower adjacent channel.

14. An apparatus for transmitting an augmentation video signal along with a standard video signal comprising:
a carrier generator for generating a carrier signal having in-phase and quadrature phase signal components at a frequency of a standard television channel;
modulator means for modulating said in-phase component with said standard video signal and said quadrature phase component with said augmentation video signal;
means for suppressing an upper sideband produced by modulating said quadrature component, whereby only a lower sideband produced by modulating said quadrature component remains: and,
means operative during a blanking interval of said video signal for suppressing all quadrature related modulation components, whereby only in-phase signal components are produced during said blanking interval.

15. The apparatus for transmitting according to claim 14, further comprising means for alternately switching the phase of said quadrature components, whereby the quadrature signal for adjacent video lines have opposite phases.

16. The apparatus of claim 14 wherein said means for suppressing said upper sideband is a reverse Nyquist filter having a linearly decreasing amplitude versus frequency response in the vicinity of said carrier signal frequency.

17. The apparatus of claim 14 comprising means for suppressing an unmodulated quadrature component whereby only a remaining lower sideband quadrature remains.

18. The apparatus of claim 15, wherein said in-phase and quadrature components are alternately switched only during portions of said augmentation signal.

19. The apparatus of claim 16 wherein said reverse Nyquist filter linearly decreasing amplitude versus frequency response begins at a frequency which substantially coincides with said television channel lower edge.

20. In a system for producing high definition television signals which include a standard video signal transmitted as an upper sideband, and an augmentation video signal, transmitted as a lower sideband of a single carrier frequency, a receiving apparatus for recovering said augmentation video signal comprising:
means for converting a received carrier frequency signal containing an upper and lower sideband into an intermediate frequency signal which contains an unmodulated component and in-phase and quadrature phase components;
means for generating a local carrier signal having a frequency and phase locked to the frequency and phase of said unmodulated component contained in said intermediate frequency signal;
means for phase shifting said local carrier signal to obtain a substantially 90° phase relationship with quadrature modulation components contained in said intermediate frequency signal; and
means for phase demodulating said quadrature components using said local carrier.

21. The receiving apparatus of claim 20 wherein said means for generating said local carrier signal comprises:

means for gating said intermediate frequency signal during an interval when said intermediate frequency signal is known to contain only in-phase components; and,
a phase locked loop connected to receive a signal from said means for gating, said phase locked loop including a voltage controlled oscillator becomes phase locked during said interval with a component of said intermediate frequency signal.

22. The receiving apparatus of claim 20 wherein said means for phase shifting comprises:
means for dividing said local carrier signal into first and second quadrature related components;
first and second modulators for receiving said first and second quadrature related components and controlling the amplitude of said components;
means for combining said amplitude controlled quadrature related components;
first and second phase detectors connected to receive said combined related components and said intermediate frequency signal;
first and second sampling means connected to each of said phase detectors for sampling and holding a voltage from said phase detector representing the phase difference between said intermediate frequency signal and said third and fourth components, and applying first and second control voltages to said first and second multipliers for maintaining said second and third quadrature components in a predetermined phase relationship with said intermediate frequency signal.

23. The receiving apparatus of claim 22, wherein said first and second sampling means sample said phase detector output at a time when said received carrier has a known phase.

24.The receiving apparatus of claim 23 wherein said time of sampling occurs during a blanking interval of said standard video signal.

25. A method for transmitting an augmentation video signal for increasing the width and resolution of a standard video signal comprising:
generating in-phase and quadrature phase components of a carrier signal of a television channel;
modulating said in-phase component of said carrier signal with a standard NTSC video signal;
modulating said quadrature component of said carrier signal with said augmentation video signal, whereby upper and lower sideband components of said quadrature component are produced;
combining at least one of said quadrature components containing said augmentation video signal with said in-phase modulation components, whereby said modulated carrier signal is produced having at least one quadrature component containing said augmentation signal; and,
alternating the phase of said combined quadrature components 180° on an alternate periodic basis, whereby the effects of a DC component in said augmentation signal on said carrier signal phase are reduced.

26. The method of claim 25 further comprising suppressing said quadrature carrier component as well as a sideband component produced by said augmentation signal. :

27. The method of claim 26 wherein said sideband is suppressed using a filter having a Nyquist response.

28. The method of claim 12 wherein said filter has an amplitude versus frequency response which decreases linearly beginning at a frequency at an upper edge of said lower adjacent channel.

29. The method of claim 25 further comprising:
receiving said modulated carrier signal at a distant receiving location;
converting said modulated carrier signal into an intermediate frequency signal; and
quadrature demodulating said carrier signal producing with a local carrier signal an in phase standard NTSC video signal, and a quadrature phase augmentation signal.

30. The method of claim 29 further comprising:
providing an augmentation signal demodulating carrier signal from said local carrier signal; and,
alternately phase shifting said demodulating carrier on a line by line basis.

31. The method of claim 29 further comprising establishing the phase of said local carrier frequency during a blanking interval of said standard NTSC signal.

FIG. I

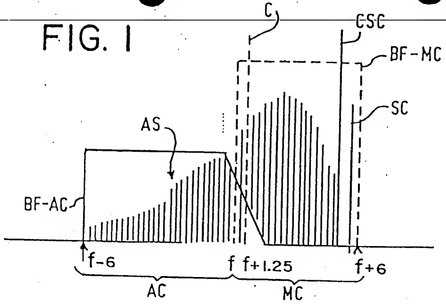


FIG. 2A

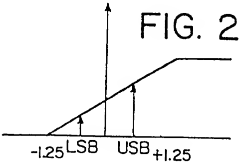


FIG. 2B

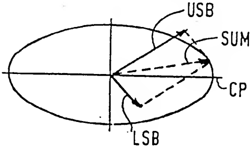


FIG. 2C

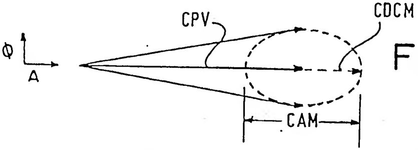
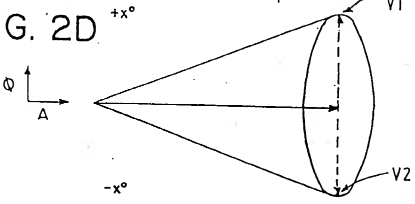


FIG. 2D



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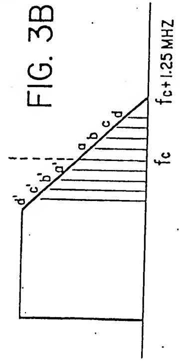

FIG. 3B

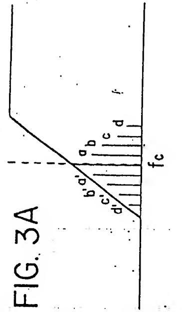

FIG. 3A

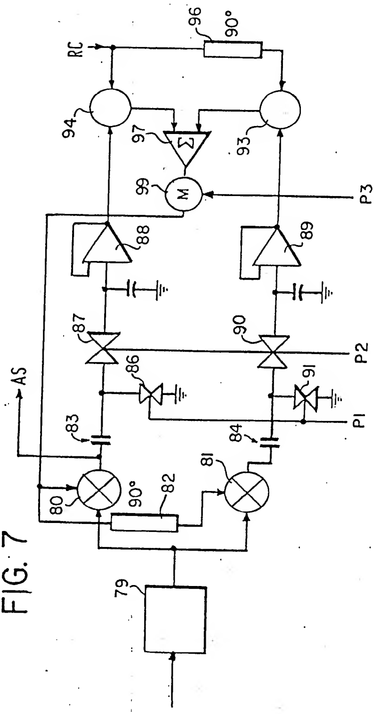

FIG. 7

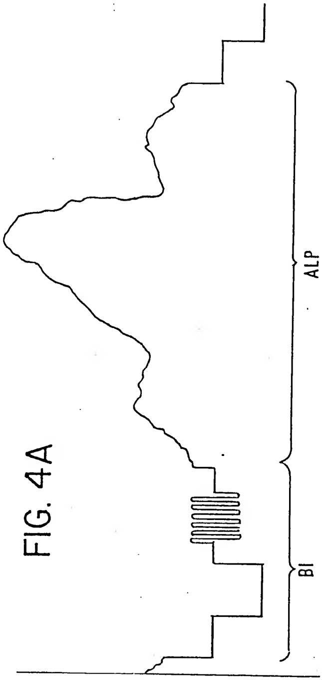

FIG. 4A

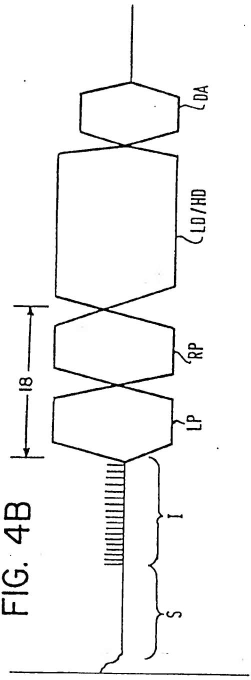

FIG. 4B

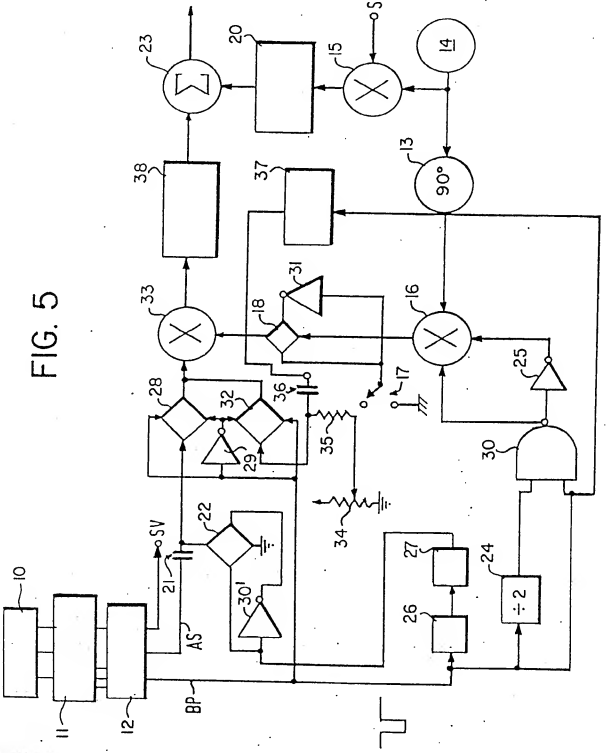
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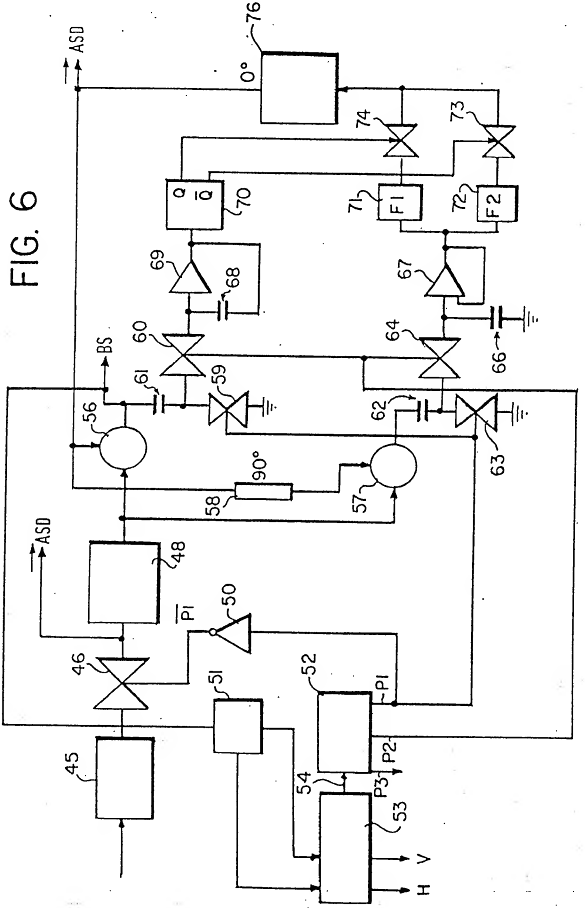
FIG. 5

FIG. 6



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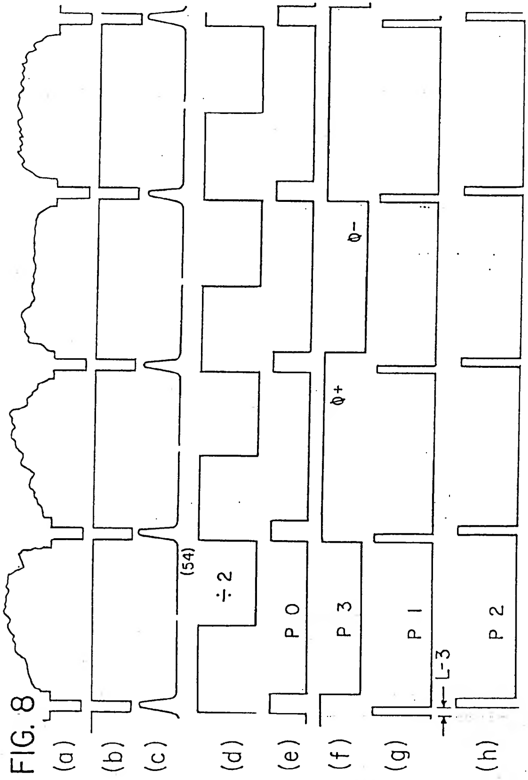

FIG. 8

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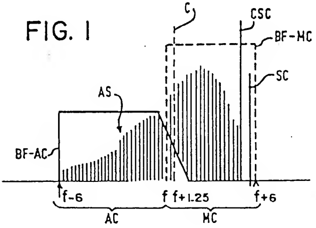
(54) System for broadcasting HDTV images over standard television frequency channels.

(57) Method and apparatus for transmitting and receiving an HDTV signal over standard television bandwidth channels. The system provides for quadrature modulation of the standard NTSC carrier with an augmentation signal containing components of the HDTV signal. The upper sideband portion of the quadrature modulation is suppressed along with the carrier produced from the quadrature modulation. The resulting lower sideband extends into the lower adjacent channel frequency spectrum, and is transmitted along with the standard NTSC video signal. Improved compatibility is achieved with existing NTSC signals by varying on an alternate line basis the phase of the augmentation signal to avoid the consequence of carrier phase shift from DC components in the quadrature modulated signal. A time gated demodulator is provided at the receiver for accurately tracking the phase of the carrier, permitting accurate demodulation of the quadrature augmentation signal. A demodulation circuit is described having phase shift compensation for removing the effects of phase delays incurred during processing of the received broadcast signal.

FIG. I



EP 0 311 188 A3

European Patent Office

EUROPEAN SEARCH REPORT

Application Number

EP 88 20 2130

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.4)
A	US-A-4 535 352 (HASKELL) * Column 4, line 24 - column 7, line 36; figures 1-8 *	1,5	H 04 N 11/00 H 04 N 7/00
A,D	IEEE INTERNATIONAL CONFERENCE ON CONSUMER ELECTRONICS, 2nd-5th June 1987, pages 80-81, New York, US; Y. YASUMOTO et al.: "New extended definition tv using quadrature modulation of picture carrier with reverse nyquist filter"	1,5,9, 14,20, 25	
A	US-A-4 631 574 (J.L. LOCICERO et al.)		
			TECHNICAL FIELDS SEARCHED (Int. Cl.4)
			H 04 N

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	10-05-1989	YVONNET J.W.

CATEGORY OF CITED DOCUMENTS

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(19) Europäisches Patentamt
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(11) Veröffentlichungsnummer: **0 329 158 A2**

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(51) Int. Cl.4: H04L 25/48 , H04L 27/00

Ein Antrag gemäss Regel 88 EPÜ auf Berichtigung einer zusätzlichen Seite der Beschreibung liegt vor. Über diesen Antrag wird im Laufe des Verfahrens vor der Prüfungsabteilung eine Entscheidung getroffen werden (Richtlinien für die Prüfung im EPA, A-V, 2.2).

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(84) Benannte Vertragsstaaten:
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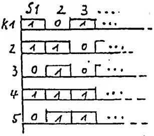
(54) **Verfahren für die digitale und/oder analoge Codierung von Information eines, zweier oder mehrerer Kanäle und/oder Frequenz- oder Bandbreitenreduzierung und/oder Erhöhung der Übertragungssicherheit.**

(57) Diesbezüglich ist bisher bekannt eine frequenz- oder zeitmultiplexe Zusammenfassung von Kanälen. Allerdings ist hierfür ein grosser Aufwand und eine grosse Bandbreite erforderlich. Bei der Erfindung werden die seriell angeordneten Codeelemente einzeln parallel geordnet und alle zusammen zu einem Codewort vereinigt. Eine Übertragungssicherheit wird in der Weise erreicht, indem die Information in PDM-Pulse umgewandelt wird und diese Impulse in die Periodendauern von Halbperioden bezw. Periodendauern umcodiert , die dann in einer ununterbrochenen Folge von positiven und negativen Halbperioden gesendet werden.

Fig.20



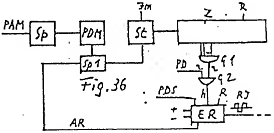


Fig.36

EP 0 329 158 A2

Verfahren für die digitale und/oder analoge Codierung von Information eines, zweier oder mehrerer Kanäle und/oder Frequenz oder Bandbreitenreduzierung und/oder Erhöhung der Übertragungssicherheit.

Die vorliegende Erfindung befasst sich mit einem Verfahren für die digitale und/oder analoge Codierung von Information eines, zweier oder mehrerer Kanäle und/oder Frequenz-oder Bandbreitenreduzierung und/oder Erhöhung der Übertragungssicherheit.

Für die Übertragung von Information mehrerer Kanäle über einen Weg sind bisher frequenz- und zeitmultiplexe Verfahren wie z.B. die Trägerfreqeunztechnik und die Pulscodemodulation bekannt. Ein Nachteil dieser Verfahren ist, dass sie grosse Bandbreiten und einen grossen Aufwand benötigen.

Aufgabe der vorliegenden Erfindung ist es die Information eines, zweier oder mehrerer Kanäle mit weniger Bandbreite zu übertragen und die Information zweier oder mehrerer Kanäle über einen Kanal mit weniger Bandbreite als für die Summe der Einzelkanäle erforderlich wäre, zu übertragen. Dies erfolgt in der Weise, indem die synchron bezw. quasisynchron angeordneten Codeelemente der verschiedenen Kanäle parallel geordnet werden und alle zusammen zu einem Codewort vereinigt und übertragen werden.Ausserdem soll noch die Übertragungssicherheit erhöht werden. Dies erfolgt in der Weise, indem die PAM-Impulse in PDM, PPM und PFM-Impulse in sinusförmige Halbperioden bezw. Periodenimpulse bezw. Codeelemente umgewandelt werden, die in einer ununterbrochenen Folge von positiven und negativen Halbperioden gesendet werden. Die Halbperiodendauer bezw. Periodendauer ist dabei ein Mass für die PDM-PPM und PFM- Impulse.

Die Erfindung kann z.B. angewendet werden zum Zusammenfassen von Telex, Teletex, Telefax, digitalen Fernsprech- Datenkanälen. Auch bei Gemeinschaftsanschlüssen und Wählsternschaltern kann die Erfindung vorteilhaft eingesetzt werden.

Weiterhin zeigt die Erfindung Möglichkeiten von vorteilhaften Codierungen neuer Fernsehtechniken zur Verbesserung von C-MAC D-MAC, D2-MAC usw. Weiterhin kann sie auch eingesetzt werden bei der Weiterentwicklung des HDTV-Verfahrens. Alle diese neuen Fernsehverfahren sind durch einen Bandbreiten mangel in ihren Möglichkeiten sehr eingeengt.

Nachstehend wird die Erfindung an Hand von Zeichnungen näher erläutert. Diese stellen dar:

Fig.1 Prinzip einer codemultiplexen Anordnung

Fig.2 Bisherige Erezeugung von Phasensprüngen z.B. bei der 4 PSK

Fig.3 bis 8 Erzeugung von Phasensprüngen

Fig.9 Erzeugung von Amplitudenstufen

Fig.10,11 und 13 Darstellung einer doppelten QAM und Vektordiagramm einer höherwertigen Codierung

Fig.14 Vektordiagramm einer doppelten QAM

Fig.16 Anordnung der Codierpunkte bei einer mehrwertigen Codierung mittels Amplitudengrössen und Phasenlage

Fig.15 Übersicht für die Erzeugungvon Phasen-und Amplitudenstufen

Fig.17 Erzeugung von Phasensprüngen

Fig.18,19,20,21,24,28 Codemultiplexe Beispiele

Fig.22,23 Übersicht eines Fernsehsenders und Empfängers

Fig.25,26,27 Duplexverkehr über Leitungen und Funk mit nur einem Wechselstrom mit Phasennachstellung

Fig.29 Kompensierung von Überlappungen

Fig.30,31,32 Erzeugung und Umsetzung von PDM-Impulsen in Halbperiodenimpulse

Fig. 33 bis 38 Erzeugung und Umsetzung von PDM- Impulse in einen Wechselstrom

Fig.39 bis 44 Codierungen gemäss der Erfindung für das Fernsehen

Fig. 45,46,62,63 Doppelbinäre und Doppelduobinäre Anordnung von Codeelementen

Fig.47,48,49 Schaltungsübersichten für das Fernsehen

Fig. 50 bis 55 Codierungen von Farbfernsehsignalen

Fig.56,57,58 Mehrfachausnützung von Übertragungswegen PDM-codierter Signale

Fig.59,60 Auswertung von phasenmodulierten Signalen

Fig.64 Schaubild über Abhängigkeit der frequenzmodulierten Schwingung von der Amplitude und Frequenz der Modulationsschwingung

Eine einfache Art Phasensprünge zu realisieren ist in den Fig.3,4,5,6 und 7 bechrieben.Zuerst wird an Hand der Fig.3 dies näher erläutert. Auf der Sendeseite S werden Rechteckimpulse mit einer Frequenz von 1 MHz angeschaltet. Wird, wie in der Fig.3c dargestellt, in den Übertragungsweg ein Tiefpass TP 5,5 MHz eingeschaltet, erhält man beim Empfänger E beinahe noch einen Rechteckimpuls . Wird, wie in der Fig. 3b eingezeichnet, ein Tiefpass TP von 3.5 MHz eingeschaltet, ist die senkrechte Flankensteilheit nicht mehr vorhanden, wird dagegen wie in der Fig 3a dargestellt, der Tiefpass auf 1,5 MHz reduziert, so erhält man beim

Emplänger E einen sinusähnlichen Wechselstrom mit der Periodendauer der Rechteckperiode. Da sich also die Periodendauer gegenüber dem Rechteckimpuls nicht ändert, kann man durch Veränderung der Periodendauern der Rechteckimpulse auch die Phase bezw. Frequenz des in der Fig 3a dargestellten sinusförmigen Wechselstromes ändern. Da eine solche Änderung immer beim Nulldurchgang erfolgt, erfolgt eine kontinuierliche Änderung und werden kaum Oberwellen erzeugt, d.h. die Übertragung ist schmalbandiger als bei den bisher üblichen Phasentastungen. In der Empfangsstelle kann dann auch die Änderung der Periodendauer als Mass für den Phasensprung vorgesehen werden. Eine solche Auswerteschaltung wird noch später beschrieben.

In der Fig4 sind Rechteckimpulse mit verschiedenen Periodendauern T =f, T=f1 und T=f2 dargestellt. Nach einer analogen Anordnung nach der Fig 3a würde man auf der Empfangsseite einen sinuslörmigen Wechselstrom mit den Periodendauern T=1/f,T=1/f1 T=1/f2 erhalten. Da bei Phasensprüngen sich die Frequenz des Wechselstromes sich verkleinert oder vergrössert, entspricht die Frequenzänderung einem Phasensprung. Aus der Fig.2, die eine Phasentastung herkömlicher Art darstellt, geht dies deutlich hervor. Man sieht in dieser, dass bei jeder Phasenänderung eine Frequenzänderung erfolgt, jedoch nicht in kontinuierlicher Weise. Daher ist es auch schwer aus der Periodendauer auf der Empfangsseite die Grösse des Phasensprungs zu ermitteln.

Um die Frequenzänderungen und damit auch das Frequenzband klein zu halten, kann man jeden Phasensprung in Stufen zerlegen. In der Fig 5 ist schematisch dies aufgezeichnet. In dieser ist T/2 die Halbperiodendauer eines Impulses und entspricht 180 Grad. Dieser Winkel wird in 36 Stufen zu je 5 Grad eingeteilt. Soll ein Phasensprung von 40 Grad zustandekommen, so wird die Halbperiode T/2 4 mal um 5 Grad gekürzt und natürlich die andere Halbperiode ebenfalls . Die Halbperiodendauer gegenüber dem Bezugsimpuls ist dann T1/2. Nach dem Phasensprung kann man entweder diese Frequenz belassen, oder aber wieder auf die Frequenz T/2 umschalten, indem man einen Phasensprung von 5 Grad in entgegengesetzter Richtung vorsieht. Gegenüber der Bezugsphase wäre dann immer noch eine Phasenverschiebung von 30 Grad vorhanden. In der Fig.6 sind zeitlich 4 mal die Perioden der Bezugsphase und 4 mal die Perioden der um 2x5 Grad gekürzten Perioden eingezeichnet. Beim Vergleich nach der 4. Periode ist der Unterschied von 40 Grad gegenüber der Bezugsphase ersichtlich.

In der Fig 7 ist eine Schaltung einer Ausführungslorm der Erfindung dargestellt. Es wird angenommen die Periodendauer in 72 Stufen zu unterteilen und zwar mit Phasensprungstufen von 5 Grad. Jeder Stufe sollen 10 Messimpulse zugeordnet werden, so sind für die Periodendauer 72x10=720 Messimpulse und für die Halbperiodendauer 360 Messimpulse erforderlich. Auf der Sendeseite brauchen immer nur die Halbperioden codiert werden. Die 2. Halbperiode wird dann jeweils über den Codierer Cod gesteuert. Werden Phasensprungstufen von 5 Grad vorgesehen, so sind für die Halbperiode, wenn die Änderung voreilend sein soll, 350 und bei einer nacheilenden Phasenänderung 370 Messimpulse erforderlich. Das Zählglied Z in der Fig 7 muss also mindestens 370 Ausgänge haben. Die Massimpulsfrequenz hängt also von der Codierfrequenz ab. Im Beispiel der Fig 7 wird im Oszillator Osc der Steuerwechselstrom für die Messimpulse erzeugt. Man kann damit unmittelbar über das Gatter G1 das Zählglied steuern, oder aber auch Pulse mittels eines Schmitt-Triggers oder einer anderen Schaltung erzeugen und mit diesen Pulsen dann das Zählglied Z schalten. Man kann auch durch Veränderung der Oszillatorfrequenz die Impulsdauer ändern. Angenommen wird der Ausgang Z2 am Zählglied Z markiert 370 Messimpulse, also die nacheilende Phasenverschiebung, dann wird vom Codierer Cod über g2 ein solches Potential an den einen Eingang des Gatters G2 gelegt, dass dann beim Erreichen des Zählgliedes Ausgang Z2, über das dann z.B. dasselbe Potential an den anderen Eingang von G2 gelegt wird, dass sich das Potential am Ausgang von G2 ändert , z.B. von h auf l. Im elektronischen Relais ER hat dies zur Folge, das Pluspotential + an den Ausgang J gelegt wird. Über die Verbindung A ist der Codierer Cod mit dem elektronischen Relais Er verbunden. Beim nächsten Überlauf des Zählgliedes Z bis Z2 wird über die Verbindung A ER so gesteuert, dass an den Ausgang J minus Potential - angelegt wird. Am Ausgang von ER können also bipolare Rechteckimpulse abgenommen werden. Man könnte genau so unipolare Rechteckimpulse erzeugen. Dieser Vorgang wiederholt sich, solange vom Codierer Cod Potential an G2 angelegt wird. Sind z.B. 5 Phasenstufen für einen Phasensprung vorgesehen, so wird das Zählglied Z 10mal bis Z2 geschaltet. Beim Ausgang Z2 erfolgt die Rückschaltung des Zählgliedes über das Gatter G4, R. Es können also durch eine verschieden grosse Zahl von Ausgängen am Zählglied Z und/oder durch Veränderung der Oszillatorfrequenz die Impulsdauer, die Stufenzahl und die Grösse der Stufen eingestellt werden. Die Steuerung dieser Varianten erfolgt über den Codierer Cod. Über fA kann eine Umschaltung der Oszillatorfrequenz, über die Anschlüsse g2,g3,... der Stufenzahl und ggf. der Phasenwinkeländerung und der Stufengrösse und über A die Amplituden der Rechteckimpulse J erfolgen. Im Beispiel sind 2

Grössen + (A) +, -(A)- vorgesehen. Die Rechteckimpulse J werden dann an einen Tiefpass analog der Fig 3 geschaltet und über einen Übertrager Ü z.B. auf den Übertragungsweg ggf. unter Zwischenschaltung eines Filters Fi. gegeben.

Am Gatter G1 muss über B noch Beginnpotential angelegt werden damit die Oszillatorpulse zur Wirkung kommen. Mit dieser Anordnung sind also folgende Codierungen möglich: eine voreilende, eine nacheilende , keine Phasenverschiebung. Diese können dabei auch stufenweise erfolgen. Die Phasendifferenz oder die Bezugsphase kann verwendet werden. Zusätzlich kann eine Amplitudencodierung ggf. stufenweise vorgesehen werden. Eine weitere Möglichkeit besteht darin die Codierung beim positiven oder negativen Impuls bezw. Halbwelle vorzunehmen. Auch die Zahl der Rechteckimpulse ist ein weiteres Codemittel.

Man kann auch eine Harmonische der Rechteckimpulse aussieben. Erfolgt dies z.B. bei der 3. Harmonischen.so sind 3 Perioden in einem plus.minus-Impuls enthalten. In diesen 3 Periodendauern sind dann auch, wenn die Impulsdauer verändert wird, die Phasenverschiebungen enthalten.

In den verschiedensten Schaltungen , wie z.B. bei der Quadraturamplitudenmodulation (QAM) werden um 90 Grad gegeneinander phasenverschobene Wechselströme benötigt. In der Fig.8 ist ein Schaltungsprinzip zur Erzeugung solcher phasenverschobener Wechselströme gleicher Frequenz dargestellt. Analog der Fig.7 wird das Zählglied Z durch einen Wechselstrom, der im Oszillator Osz erzeugt wird und über das Gatter G , an dessen anderen Eingang ein Beginnpotential B liegt,geführt wird, gesteuert. Im Beispiel sollen 4 Rechteckimpulse erzeugt werden, die gegeneinander um 90 Grad phasenverschoben sind.Hat das Zählglied Z 100 Ausgänge , so sind beim 25.,50.,75. und 100. Ausgang elektronische Relais ER1 bis ER4 analog dem ER-Relais in der Fig.7 anzuschalten.Mit diesen elektronischen Relais werden dann wie bereits in der Fig.7 beschrieben, Rechteckimpulse erzeugt. Hier sind in den ER-Relais noch Mittel, die bei bipolaren Rechteckimpulsen immer eine Potentialumkehr vornehmen und bei unipolaren Rechteckimpulsen das Potential während eines Durchlaufs wegnehmen.Die Rechteckimpulse werden dann, in der Fig.7 mit J bezeichnet, über die Filter Fi1 bis Fi4 gesendet. Der dann entstehende Wechselstrom hat jeweils 90 Grad Phasenverschiebung gegenüber dem vom nächsten Ausgang erzeugten. An Stelle von phasenverschobenen Wechselströmen kann man durch die Ausgänge auch um 90 Grad phasenverschobene Abnahmen von z.B. PAM-Proben steuern. Am elektronischen Relais ER1 ist noch ein Filter Fi0 angeordnet das z.B. nur die 3. Oberwelle des Rechteckimpulses durchlässt, sodass man hier die 3-fache Frequenz der Rechteckimpulse erhält. Die Phasenverschiebung wird dann auf die 3. Oberwelle übertragen.

Mit der Fig.7 kann man gleichzeitig auch verschiedene Amplitudenstufen erzeugen. In der Schaltung sind nur 2 gekennzeichnet, In der Fig.9 ist eine weitere Möglichkeit verschiedene Amplitudenstufen zu erzeugen. Der z. B. in der Fig.7 erzeugte Wechselstrom wird einem Begrenzer zugeführt, in dem die Steuerimpulse erzeugt werden. Über den Anschluss Code werden die Kennzustände zugeführt, die eine Umschaltung auf die durch den Code bestimmten Amplitudengrösse vornehmen und zwar im Codierer Cod. Die Umschaltung auf eine andere Amplitudengrösse erfolgt immer beim Nulldurchgang . Die Grösse der Amplituden wird durch die Widerstände R1 bis R4 , die in Wechselstromkreisen angeordnet sind, bestimmt. Elektronische Relais I bis IVes, die durch den Codierer Cod gesteuert werden, schalten die verschiedenen Widerstände in den Wechselstromkreisen ein. Am Ausgang A erhält man dann 4 verschieden grosse Amplituden.

Es ist auch bekannt eine Information durch die Halbwellen bezw. Perioden eines Wechselstromes zu codieren, bei einem Binärcode sind dann die Kennzustände grosser und kleiner Amplitudenwert. Werden 2 solcher Codierwechselströme gleicher Frequenz um 90 Grad phasenverschoben und addiert, so können diese mit einem Wechselstrom gleicher Frequenz übertragen werden. In der Fig. 10a,b sind die Kanäle K1 und K2, die durch die Perioden als Codeelemente codiert werden mit den Kennzuständen grosser Amplitudenwert =1 und kleiner Amplitudenwert=0. Wird einer gegen den anderen um 90 Grad phasenverschoben, so können sie addiert werden. In der Fig. 11 ist ihr Vektordiagramm dargestellt. Der Kanal K1 hat den Vektor K1 (u) und der Kanal K2 den Vektor k2 (v). Die beiden Kennzustände der beiden Wechselströme sind mit u1/uo und v1/vo bezeichnet. Werden nun beide addiert, so erhält man die 4 Summenvektoren I,IV und II,III. Man sieht, dass die Vektoren II und III nicht mehr auf der 45 Grad Linie liegen. Die Auswertung ist dadurch etwas schwieriger. Für die Auswertung der Binärsignale genügen 4 Möglichkeiten, die man alle auf die 45 Grad Linie legen kann, in der Fig.11 mit (II) und (III) bezeichnet. In der Fig.13 sind die 4 Möglichkeiten dargestellt, 00,11,10,01. Sind alle 4 Möglichkeiten auf dem 45 Grad Vektor, wie in der Fig.11 dargestellt, so kann man diese durch 4 verschiedene grosse Amplituden codieren, d.h.mit einem sinusförmigen Wechselstrom. In der Fig.9 ist eine solche Möglichkeit dargestellt. Um binäre Signale von 2 Kanälen zu übertragen genügt also ein mehrwertiger quaternärer Code; wie z.B. die 4 PSK oder 4 QAM. Diese Codierungen sind auf eine Periode verteilt. In der

Fig.9 sind die positive und negative Halbwelle gleich gross, es liegt dann bei der Über tragung eine Gleichstromfreiheit vor. Man kann die positive und negative Halbwelle als zusätzliches Kriterium ausnützen. Man kann dann die 4 Amplitudenkennzustände verteilen , 2 auf die positive und 2 auf die negative Halbwelle. Diese können dieselbe Grösse haben, also z.B. in Fig.11,I +IV für die positive und negative Halbwelle. Damit dieser Codierwechselstrom immer über dem Störpegel liegt, muss der Codierwechselstrom immer eine bestimmte Grösse aufweisen, z.B. wie in Fig.11 (III). Die Amplitudengrösse IV wird man dann etwas vergrössern.

Eine Verkleinerung von z.B. binärcodierten Wechselströmen mit den Halbwellen bezw. Perioden als Codeelemente ist bereits bekannt. Voraussetzung hierfür sind Phasenverschiebungen der Probeentnahmen. Die vorliegende Erfindung zeigt eine weitere Möglichkeit auf,die Frequenz insbesonders binärcodierter Information zu verkleinern. In der Fig. 1 ist ein Kanal K mit einem Binärcode 1,0,1,1,..aufgezeichnet. Soll die Frequenz des Kanales verkleinert werden in 2 Kanäle mit der halben Frequenz, so müssen jeweils 2 seriell angeordnete. Binärwerte des Kanales K parallel auf die Kanäle Kv1 und Kv2 verteilt werden, z.B. die 4 Werte 1,0,1,1 des Kanales K der Wert 1 auf Kv1, der Wert 0 auf Kv2, der Wert 1 wieder auf Kv1 und der weitere Wert 1 auf Kv2. Einen Wert kann man dabei immer speichern, oder man kann die Werte auch zeitlich versetzt übertragen. Bei der Auswertung muss dies berücksichtigt werden. Eine gleichzeitige Übertragung von 2 Kanälen wurde bereits schon in den Fig.11 und 13 dargelegt. Wie aus der Fig. 13 ersichtlich ist, sind 4 Kombinationen möglich.

In der Fig.10 sind 4 Codierwechselströme K1-K4 mit den Codeelementen Periode und den Kennzuständen grosser und kleiner Amplitudenwert gleicher Frequenz dargestellt. Will man alle 4 auf der Basis der QAM übertragen, müssen diese folgende Phasen aufweisen, K1=0Grad, K2=90 Grad, K3=90 Grad und K4=180 Grad. K1/K2 und K3/K4 werden zu einem Codierwechselstrom entsprechend der Fig.9 zusammengefasst und addiert. In der Fig.14 ist hierfür das Vektordiagramm dargestellt. Man sieht, dass 16 Kombinationen möglich sind. Weiterhin ist hieraus ersichtlich, dass nur 4 Werte auf dem 45 Grad vektor liegen. Bei der Auswertung müssen für die anderen Werte noch die voreilende bezw. nacheilende Phasenverschiebung berücksichtigt werden. Die phasenverschobenen Wechselströme werden in einer Anordnung wie in der Fig.8 dargestellt, erzeugt und 2 Anordnungen nach der Fig.9 zugeführt, wobei diese Wechselströme gegeneinander um 90 Grad phasenverschoben sind.

Man kann auch einen Summenwechselstrom und einfachen Codierwechselstrom addieren, Voraussetzung ist eine 90 Grad Phasenverschiebung gegeneinander. Dabei entstehen 8 Kombinationsmöglichkeiten.

Auch 4 Kanäle können Codiermultiplex , wie in der Fig. 1 dargestellt, übertragen werden. Einmal sind 16 Kombinationen notwendig. Man kan hierfür auch bekannte Codierungen vorsehen, wie z.B. die 16 PSK, die 16 QAM die 8 PSK. Zur Codierung ist hier jeweils eine Periode erforderlich, wenn Phasenverschiebungen gemäss der vorliegenden Erfindung vorgesehen werden. An Stelle der doch eng zusammenliegenden Kennzustände bei der doppelten QAM nach Fig. 14, kann man auch eine beliebige Codierung vornehmen. In Fig.16 wird die Codierung durch 30 Grad Phasenunterschiede und durch 3 und4 Amplitudenstufen vorgenommen. Falls man noch grössere Sicherheit haben will, kann man die 4 Amplitudenstufen BPh noch aufteilen. Auf der Nullinie können noch Stufen untergebracht werden. Man kann also jede Halbwelle für eine solche Codierung vorsehen. Will man jedoch eine Übertragung über drahtgebundene Übertragungswege vornehmen, ist es zweckmässig die negative Halbwelle mit derselben Codierung zu übertragen, damit man eine Gleichstromfreiheit hat. Mit derselben Methode kann man auch eine Verkleinerung vornehmen. In Fig.1 soll der Kanal nur mit der viertelchen Frequenz übertragen werden. Jeweils 4 seriell angeordnete Binärelemente 1 und 0 werden parallel wie in der Fig. 1 a,b vorgesehen, angeordnet. Die Werte 1,0, 1,1 des Kanales K werden dann parallel aufgeteilt auf den Kanal Kv1 "1", Kanal Kv2 "0", Kanal Kv3 "1" und Kanal Kv4 "1". Im Codierer wird dann für die jeweilige Kombination der vorbestimmte Codierpunkt ermittelt und auf die Phase und Amplitude des Codierwechselstromes übertragen. Die Phase wird in der Fig.7 festgelegt, ggf. kann man mit dieser auch gleich die Am plitude codieren, und in der Fig.9 kann man dann die erforderlichen Amplituden codieren. In der Fig. 15 ist die Übersicht hierfür dargestellt. Im Codierer Cod erfolgt die Festlegung des Codierpunktes aufgrund der Viererkombination. Der Phasencodierer erzeugt die Halbwellen bezw. Perioden mit entsprecher Phase und der Amplitudencodierer erzeugt die dazugehörigen Amplituden. Ein Phasencodierer kann analog der Fig.7 und ein Amplitudencodierer analog der Fig.9 aussehen.

Ein Phasensprung bedeutet immer eine Änderung der Periodendauer. Diese Änderung, also Frequenzänderung, kann bei keiner weiteren Phasenänderung beibehalten werden, oder man kann bei der nächsten Periode bezw. Halbperiode wieder auf die ursprüngliche Frequenz umschalten. Da im letzteren Fall der Wechselstrom eine andere Phase aufweist, ist bei der Auswertung eine Bezugsphase erforderlich. Wie aus der Fig.4 hervorgeht kann mit

Hilfe der Schaltung der Fig.7 jede beliebige Phase beibehalten, d.h. die Frequenz beibehalten werden, die bei der Phasenänderung entstanden ist. Die Phasenänderungen werden immer im vorliegenden Fall beim Nulldurchgang vorgenommen. In der Fig.16 kann man eine Bezugsphase BPh vorsehen, von der aus vor- und nacheilend 2x 30 Grad eine Phasenverschiebung vorgenommen wird.

In der Fig. 17 ist eine Erzeugung der Phasensprünge der Fig. 16 nach dem Prinzip der Fig.7 dargestellt. Der Winkel von 360 Grad wird durch 3600 Pulse gekennzeichnet. Liegt nur eine Amplitudenänderung mit der Bezugsphase vor, so wird das Zählglied immer von 0 bis 360 Grad durchgeschaltet. Die Steuerung erfolgt dabei über den Codierer Cod, der bereits in der Fig.7 beschrieben wurde. Die Amplitudenänderung erfolgt dabei wie in der Fig.7 oder wie in der Fig.9 dargestellt. Soll der Phasensprung Ph1 in Fig.16 erfolgen, so muss, wenn eine Gleichstromfreiheit erforderlich ist, jede Halbperiode bis zum Ausgang 195 geschaltet werden. Eine Bezugsphase ist bei der Auswertung nicht notwendig, weil, solange keine weitere Phasenänderung erfolgt, durch die Periodendauer ja die eindeutige Phase festgelegt ist. Liegt die Codierung auf dem Vektor Ph3, so ist die Periodendauer 330 Grad, d.h. beim Ausgang 165 erfolgt immer eine Umschaltung . Die Phasenverschiebung ist hierbei immer auf die Periodendauer bezogen. Würde z.B. im letzten Fall die Phasenverschiebung auf die Halbperiode bezogen, so müsste jeweils eine Rückschaltung beim Ausgang 150 erfolgen. Andere Methoden der Erzeugung von Phasensprüngen können genau so verwendet werden.

Die Auswertung der Phasensprünge erfolgt in bekannterweise durch Abmessung der Periodendauern mittels einer überhöhten Steuergeschwindigkeit von Zählgliedern, z.B. in der europäischen Patentanmeldung 86104693.6 offenbart.

Bei der Auswertung der Fig. 14 ist eine Bezugsphase erforderlich. Die Amplitudenpunkte 1 bis 4 sind unmittelbar auf der Bezugsphasenlage, während die anderen 12 Codierpunkte voreilend und nacheilend zur Bezugsphase angeordnet sind. Es wird angenommen die Signale sind die eines Fernsehsystems. In der Austastzeit wird dann die Bezugsphase ermittelt und zugleich Steuersignale übertragen. Dabei werden nur die Amplitudenwerte auf der Bezugsphase verwendet. Vom Übertragungsweg ÜW werden die Signale dem Eingangssatz EST zugeführt (Fig.12). Einmal gehen sie dann zu einem Begrenzer B und einmal zu einer Codeauswertung CA. Im Begrenzer werden die positiven und negativen Halbwellen zu Jp und Jn- Impulsen umgewandelt. In der Vergleichseinrichtung VE wird nun die Phase der von dem Übertragungsweg kommenden Impulse mit einem Bezugsphasenimpuls JBn verglichen. In der Fig. 12asind die vor-nacheilenden und der Bezugsphasenimpuls Jv,Jn,JB dargestellt, die mit dem aus einer Codierung ermittelte Bezugsphasenimpuls JBn verglichen werden. Die 3 möglichen Phasenwerte vor-nacheilend oder Bezugsphase werden jeweils zur Codeauswertung gegeben. In dieser werden die Amplitudenwerte ermittelt und in Verbindung mit der vor-nacheilenden oder Bezugsphase werden dann die Codierungspunkte ermittelt und über S zur weiteren Verwertung weitergesendet. Die Codierung der Bezugsphase in der Austastzeit kann z.B. so aussehen, dass man 4 mal den Punkt 2 und 4 mal den Punkt 4 auf der Bezugsphase sendet. Die Auswertung derselben erfolgt in der Bezugsphasenauswertung BA. Von dieser wird dann ein Bezugsphasenimpuls JBn zur Vergleichseinrichtung gegeben.

In der Fig.18 ist ein weiteres Ausführungsbeispiel der Erfindung dargestellt. Die 5 Kanäle K1 bis K5 sollen codemultiplex nur über einen Kanal bezw. Weg übertragen werden. Die z.B. binärcodierte Information dieser 5 Kanäle wird zuerst im Speicher Sp gespeichert. In der Fig.20 sind z.B. die Schritte der Binärzeichen dargestellt und zwar bereits synchronisiert. Zu codieren sind also jeweils 5 parallel angeordnete Schritte bezw. Impulse S1,2,3,... Die Schritte von S1 sind 1-1-0-1-0. Für die Codierung dieser 32 Kombinationen sind 5 bit erforderlich. Im Beispiel werden diese mit den Amplituden der Halbwellen eines Wechselstromes mit den Kennzuständen grosser und kleiner Amplitudenwert und mit einem voreilenden und einem nacheilenden Phasensprung von 36 Grad codiert, wie in der Fig.19 gezeigt ist. Vom Speicher Sp der Fig.18 werden die Binärwerte dem Codierer Cod zugeführt und in diesem in einen entsprechenden Code umgewandelt. Im Decodierer der Empfangsseite werden entsprechend dem Code den 5 Kanälen die entsprechenden Schritte wieder zugeordnet.

In der Fig.21 ist eine weitere Anwendung der Erfindung für die Codierung und Übertragung der Signale beim Farbfernsehen dargestellt. Das Luminanzsignal wird mit 6 MHz abgegriffen. Dieses Prinzip ist bereits schon in der Offenlegungsschrift P 32 23 312 offenbart. Die Farben rot und blau sollen je mit 1,2 Mhz abgegriffen werden, d.h. auf 5 Luminanzabgriffe trifft je ein Rot-und Blauabgriff. Die Luminanzabgriffe sind mit I,II, III,IV,V bezeichnet. Diese Probeentnahmen werden mit 8 bit codiert, im Beispiel binärcodiert. Mit dem Abgriff III müssen dann auch die Abgriffe für rot und blau erfolgen. Die Probeentnahmen von rot und blau werden im Beispiel mit 6 bit binärcodiert. Während der Übertragung der 5 Luminanzprobeentnahmen wird auch gleichzeitig der Code für die Farbprobeentnahmen rot und blau gesendet. Mit dem Abgriff von rot und blau könnte man mit der Übertragung der Farbe und mit der Probeentnahme I des Lumi-

nanzsignales beginnen. Man kann auch alle 5 Luminanzprobeentnahmen und Farbsignalproben speichern und erst nach der 5. Probeentnahme mit der Übertragung aller Fernsehsignale beginnen. In der Fig.21a sind die binären Codes aller zu übertragenden Signale aufgezeichnet. Die 8 bit 1-8 der Luminanzprobeentnahmen sind jeweils parallel angeordnet. Seriell sind dann unter 9,10 digitale Ton-und sonstige Signale T + So , die 6 bits des Rotsignales und nochmals die Ton-und sonstigen Signale und unter 11,12 wieder die Ton-und sonstigen Signale und die 6 bits des Blausignals angeordnet. Zweckmässig ist es, wenn man die Luminanzproben I bis V beim Sender noch speichert und die Farbcodes für rot und blau mit den vorhergehenden Luminanzproben sendet, sodass dann beim Empfänger sich eine Speicherung der 5 Luminanzproben erübrigt. Es müssen dann lediglich die Rot-und Blau proben gespeichert werden. Die Ton-und sonstigen Signale müssen ebenfalls gespeichert werden und dann zeitgleich mit dem Bild dem Lautsprecher zugeführt werden. Diese Signale können natürlich auch in die Austastzeit gelegt werden. Im Beispiel sind also 12 bit für die Übertragung einer Luminanzprobe für die Ton-und sonstigen Signalproben und für die Farbprobeentnahmen erforderlich. In der Fig.21b ist ein Beispiel für die Codierung dieser 12 bits dargestellt. 5 Halb-Perioden eines Wechselstromes werden hierfür vorgesehen. Der Binärcode besteht dabei aus Codeelementen der Halbwellen mit den Kennzuständen grosser und kleiner Amplitudenwert. Zusätzlich wird noch eine voreilende und nacheilende Phasenverschiebung von 36 Grad vorgesehen, sodass man damit 12 bit erhält.

In der Fig.22 ist eine Übersicht eines solchen Fernsehsenders dargestellt. Das Steuerorgan StO steuert die Fernsehkamera FK liefert auch die übrigen Steuersignale wie Austast- und Synchonisiersignale A + S. Die Rot-Grün- und Blausignale werden einmal der Y-Matrix YM und rot und blau zugleich der Farbartaufbereitung FA zugeführt. Zugleich ist ein Konzentrator K vorgesehen, der das Luminanzsignal Y, die Farbsignale r + bl und die Ton- und sonstigen Signale abgreift. Beim Abgriff 3 wird über die Verbindung 3a ein Kriterium zur Farbartaufbereitung gegeben. In dieser wird ein Abgriff vom Rot-und Blausignal vorgenommen und beide Werte werden in den Kondensatoren C1 und C2 gespeichert. Der FA wird noch von der Y-Matrix ein Y-wert der beim 3. Abgriff vorhanden ist, zugeführt, sodass man am Abgriff 6a und 6b die Farbdifferenzsignale r-y und b-y er hält. - Man kann auch nur die Farbauszugssignale abgreifen.-Über den Baustein TSo werden die Ton- und sonstigen Signale analog über 6c und 6d dem Konzentrator zugeführt. Vom Konzentrator aus werden alle Werte einem Speicher Sp zugeführt. Vom Speicher aus werden die Signale zeitgerecht z.B. wie in Fig.21a beschrieben, einem Analog/Digitalwandler zugeführt. In diesem erfolgt eine Codierung entsprechend der Fig.21b. Während der Austastzeit erfolgt eine Umschaltung auf den Konzentrator K1 über U. Als Austastkriterium kann man z.B. einigemale das Codewort mit nur Nullen senden. - --- Auch können in der Austastzeit noch sonstige Signale So gesendet werden. Auch den Beginn einer Zeile kann man durch einen Nullcode markieren. Während der Zeile ist durch die Folge und der Zahl der Halbwellen eine Synchronisierung vorgegeben. Bei dem vorliegenden Code ist eine Nenn Frequenz von 15 MHz erforderlich. Will man nur einen Amplitudencode verwenden, sind 2 Wechselströme mit je 18 MHz erforderlich, die man dann um 90 Grad phasenverschieben könnte und addiert übertragen könnte. Es ist lediglich eine Frage der Wirtschaftlichkeit und Sicherheit welche Methode hier verwendet wird. Der vor- oder nacheilende Phasensprung wird im Beispiel durch die Periodendauer festgelegt. Es ist also dann keine Bezugsphase erforderlich. Natürlich können zur Verringerung der Frequenz mehrstufige Amplitudencodes oder/und Phasencodes verwendet werden. An den Eingang Ton T kann man z.B. das PAM-Signal anlegen,das dann innerhalb der 8 KHz-Zeit öfters abgegriffen wird. Es gibt hier zahlreiche Möglichkeiten den Abgriff 6c/6d auszunützen. In der Fig.23 ist eine Teilübersicht eines Fernsehempfängers dargestellt. Über die HF-Oscillator und Mischstufe und dem Verstärker V werden die Signale dem Demodulator DM zugeführt. In diesem werden z.B. die Signale wie sie in der Fig.21b dargestellt sind wieder gewonnen und dem Decodierer DC zugeführt. Die Farbsignale werden in der Folge der Matrix Ma weitergegeben. An diese auch das Y-Signal geschaltet. Am Ausgang der Matrix erhält man dann z.B. die Farbdifferenzsignale R-Y, G-Y und B-Y, die wie UY an die Fernsehröhre geführt werden.Der Decoder DC liefert dann noch die Austast - und Synchronisiersignale AS, die Ton- und sonstigen Signale.

In der Fig.24 ist ein Beispiel dargestellt, bei dem der Code für den Codemultiplex aus mehreren Wechselströmen gewonnen wird. Es stellt einen Binärcode dar bei dem die Halbwellen der Wechselströme als Codeelemente dienen und bei dem ein grosser und ein kleiner Amplitudenwert die Kennzustände bilden. Die zu übertragenden Kennzeichen bestehen aus Rechteckimpulsen der Frequenz 1000 Hz, wie in der Fig.24a dargestellt ist. Es sollen 20 Kanäle codemultiplex übertragen werden. Hierfür werden die Halbwellen der Wechselströme 1000, 1500, 2000, 2500 und 3000 Hz vorgesehen. Jedem Kanal kann man natürlich zeitmultiplex mehrere Kanäle niedrigerer Bitfrequenz zuführen. Dieselbe Bit-Zahl könnte man genau so mit 2 Wechselströmen mit 2000 Hz und nochmals

2 Wechselströmen mit 3000 Hz erreichen , wobei diese jeweils gegeneinander um 90 Grad phasenverschoben sein müssten, sodass sie bei der Übertragung addiert werden könnten. Wie am besten die Synchronisierung zwischen den einzelnen Kanälen hergestellt wird ist bereits bekannt (Unterrichtsblätter der DBP Heft4/6Jahr79), und es wird deshalb nicht weiter darauf eingegangen. Auf dieselbe Art kann man auch die digitalisierte Sprache bezw. mehrere Sprachkanäle gleichzeitig übertragen.

Bei einer Amplitudencodierung kann man mit demselben Wechselstrom Duplexbetrieb durchführen. Dazu ist es notwendig, dass der Gegencodierwechselstrom um 90 Grad phasenverschoben ist. In der Fig.25 ist dieses Prinzip dargestellt. Der Code kann dabei digital , ein Binärcode sein entsprechend dem Patent DE 30 10 938 oder aber auch analog entsprechend dem kanadischen Patent 1 214 227. Bei Halbwellen als Codeelemente ist bei digitaler Codierung die Frequenz 32 KHz und bei analoger Codierung 4 KHz. In der Fig.25 ist S1 das Mikrofon und E2 der Hörer des einen Teilnehmers und S2 und E1 des anderen Teilnehmers. In S1 ist noch ein Codierer, in dem aus der Sprache der Codierwechselstrom gewonnen wird. Von S1 geht der Codierwechselstrom über eine Gabel G, die Anschluss- bezw. Verbindungsleitung RL zur Gabel G des Gegenteilnehmers und zum Hörer E1. In diesem ist zusätzlich ein Decodierer, der aus dem Codierwechselstrom wieder die Sprache herstellt. Der Codierwechselstrom von S1 sei der Synchronisierwechselstrom. Von E1 wird dieser über einen Phasenschieber 90 Grad zu S2 abgezweigt, in dem er ggf. verstärkt wird. Spricht nun S2, so wird ein um 90 Grad phasenverschobener Codierwechselstrom über G,RL, G nach E2 gesendet, dort decodiert und dem Hörer als Sprache übermittelt. Wenn z.B. kurzzeitig gleichzeitig gesprochen wird, entsteht auf dem Übertragungsweg RL ein Additionswechselstrom. Eine Auslöschung wird nicht verursacht. Dieses Prinzip kann genau so beim Duplexverkehr bei der Datenübertragung vorgesehen werden. Weitere diesbezügliche Beispiele sind in der Offenlegungsschrift DE 3802088 offenbart.

Diese Methode kann natürlich auch bei Funk z.B. beim Richtfunk verwendet werden.In der Fig.26 ist eine diesbezügliche Übersicht aufgezeichnet. Der Sendewechselstrom wird hier zugleich als Codierwechselstrom mit vorgesehen. Vorteilhaft wird eine Vorstufenmodulation verwendet. Im Oszillator Osz1 wird der Sendewechselstrom erzeugt Im Amalog/Digitalwandler A1/D1 wird das Basissignal in einen Wechselstromdigitalcode umgewandelt.-Noch einfacher ist es als Oszillator und Codierer eine Anordnung nach der Fig.7 vorzusehen. Vom Codierer aus wird dann das elektronische Relais so gesteuert, dass am Ausgang J grosse und kleine Rechteckimpulse vorhanden sind, die dann im Tiefpass TP zu einem sinusförmigen Wechselstrom geformt werden. - Über nicht eingezeichnete Verstärker gelangt dann der Codierwechselstrom zur Endstufe E und zur Sendeantenne. In der Endstufe kann man noch einen Zweigstromkreis vorsehen, in dem die Oberwellen um 180 Grad phasenverschoben werden, die dann zur Kompensation dem Hauptstromkreis wieder zugeführt werden. Auf der Empfangsseite werden die Nutzsignale über einen festen Abstimmkreis einem Verstärker V zugeführt und dann an den Digital-Analogwandler D2/A2 weitergeschaltet. Das Analogsignal wird dann z.B. über eine Vermittlung weiter geleitet. Über den Verstärker V wird der Sendewechselstrom auch zu einem Phasenschieber von 90 Grad Ph abgezweigt und dann zum Oszillator Osz2 weitergeschaltet. Mit diesem wird der Oszillator synchronisiert. Über den Wandler A3/D3 , nicht eingezeichnete Verstärker und den Endverstärker E wird dann der Sender der entgegengesetzten Richtung betrieben. Der Empfänger E1 ist genau so wie der Empfänger E2 geschaltet, nur der Phasenschieber ist nicht erforderlich.

Ein Phasenschieber nach dem Prinzip der Fig.7 ist in der Fig. 27 dargestellt. In dieser ist zugleich ein Ausgleich für kleine Frequenzschwankungen vorgesehen. Für diesen Zweck wird ein Zählglied Z vorgesehen mit 1000 Ausgängen. Während einer Halbwelle des Sendewechselstromes durchläuft das Zählglied diese 1000 Ausgänge. Die Steuerimpulse Js werden im einem nicht eingezeichneten Oszillator erzeugt. Bei 90 Grad Phasenverschiebung trifft auf eine Halbwelle eine Phasenverschiebung von 45 Grad, das entspricht 250 Ausgängen. Die vom Verstärker V kommenden Sendewechselstromhalbwellen werden einem Begrenzer zugeführt, sodass am Ausgang desselben Rechteckimpulse Jp und Jn entstehen. Diese Impulse werden dem Steuerglied St zugeschaltet.An dieses werden noch die Steuerimpulse Js und das Beginnkennzeichen Be gelegt. Das Steuerglied ist so geschaltet, dass immer nur ganze Jp bezw. Jn-Impulse beim Zählglied wirksam werden. Hat während eines Impulses Jp das Zählglied den Ausgang 1000 erreicht, so kommt das Gatter G11 in Arbeitsstellung. Am Gatter G12 ist ein Jn-Impuls und nach dem Ende des Jp-Impulses durch die Verzögerung des monostabilen Gliedes mG4 kurzzeitig noch Potential angeschaltet. G12 wird wirksam und legt an den einen Eingang von G13 Potential, am anderen Eingang von G13 wurde bereits I - Potential von G11 aus angelegt. Am Ausgang von G13 erfolgt nun ein Potentialwechsel, der G16 am Ausgang umpolt. Dies hat zur Folge, dass G17 für das Zählglied ein Rückschaltepotential erzeugt. Auch an die Gatter G8,G9 und G10 wird solches Potential gelegt, dass die in Zusammenwirken mit den

belegten Ausgängen 1000, 999, 1001 eines der monostabilen Glieder mg1,mG2 oder mG3 steuern. Da der Jp-Impuls das Zählglied bis 1000 gesteuert hat, wurde nun das Gatter G9 und mG2 wirksam. Wird nun mit dem nächsten Jn-Impuls das Zählglied auf den Ausgang 250 gesteuert, so wird das Gatter G6 wirksam, das das elektronische Relais ER steuert, das entsprechend der Fig.7 einen Rechteckimpuls erzeugt, der im Tiefpass zu einer Halbwelle geformt wird. Für den Jn-Impuls sind für die Ausgangsmarkierung die Gatter G15 G14 und das monostabile Glied mG5 angeordnet. Das monostabile Glied mG2 hält sich z.B. bis zum Ausgang 260 . G6 geht dann wieder in die Ausgangsstellung. Das elektronische Relais bleibt bis zur nächsten Markierung des Ausganges 250 in die ser Stellung. Wird durch eine Frequenzschwankung nur der Ausgang 999 erreicht, so wird an Stelle von G9 das Gatter G8 markiert und mG1 und G5 beim Erreichen des Ausganges 249 zur Wirkung gebracht. Wird der Ausgang 1001 erreicht, so wird G10 und mG3 zur Wirkung gebracht und beim Erreichen des Ausganges 251 das Gatter G7. Solche Frequenzschwankungen werden also auch an den 90 Grad phasenverschobenen Wechselstrom weitergegeben. In der Fig.27 a ist das Steuerglied im Einzelnen dargestellt. Die Impulse Jn und auch das Beginnzeichen sind an das Gatter G3 geschaltet. Sind beide vorhanden, wird G3 wirksam und bringt das bistabile Glied bG in die Arbeitslage, das nun an das Gatter G1 Arbeitspotential legt. Erst jetzt kann der Jp-Impuls zur Wirkung kommen. Die Steuerimpulse Js gelangen nun über das Gatter G2, das lediglich ein Potentialumkehrgatter ist, an das Zählglied. Die weiteren Vorgänge am Zählglied sind bereits beschrieben.

In der Fig.27 kann die negative Halbwelle entweder durch den Jn-Impuls erzeugt werden, oder es wird der Durchlauf der positiven Halbwelle wiederholt, wobei die jeweils markierten Ausgänge gespeichert werden.

Der bei der Erfindung verwendete Code kann vorzugsweise ein Amplituden und/oder Phasencode sein, wie z.B. ein solcher in Fig. 16 dargestellt ist. Bei einem reinem amplitudencode kann man auch 2 Codewechselströme gleicher Frequenz vorsehen, wobei der eine dann bei der Übertragung um 90 Grad phasenverschoben wird und in der Folge mit dem anderen addiert wird.

Das Prinzip der Erfindung kann auch für die Übertragung digitalisierter Sprache. In der Fig.28 sind 5 Codierwechselströme mit einem Binärcode, wobei die Kennzustände ein grosser und ein kleiner Amplitudenwert der jeweiligen Halbwelle ist, dargestellt. Die Frequenzen sind dabei 8,12,16,20 und 24 KHz. Man erhält dabei 20 bit, werden zusätzlich 2 Wechselströme gleicher Frequenz, jedoch um 90 Grad phasenverschoben, vorgesehen, so erhalt man 40 bit, d.h. bei 8 bit Codewörtern, wie in der Fig.28a dargestellt, kann man damit 5 digitalisierte Sprachkanäle übertragen.

In den Fig 21 und 22 genügen je Zeile bei einer Abgriffsfrequenz von ca. 30 KHz (PAM) je Zeile 2 Ton- Abgriffe, die z.B. beim Beginn der jeweiligen Bildzeile und in der Mitte der Bildzei lo erfolgen können, der Abstand ist dann 32 µs. Jeder Abgriff wird dann im Analog/Digitalwandler A/D in einen 8 bit-Code umgewandelt und wird dann , wie in der Fig.21a dargestellt ist, mit den folgenden 5 Luminanzcodewörtern gesendet. In der Fig. 21a z.B. mit I/9,1o,11,12 und V/9,10,11,12. Die Abgriffe während der Bildwechselzeit müssen z.B. durch eine Zeitmessung ermittelt werden. Die Codierung erfolgt dann auch in der Bildwechselzeit.

Für das Codemultiplex kann natürlich jeder beliebige Code verwendet werden wie der AMI- oder HDH-3 Code. In den Beispielen wird vielfach ein Amplitudencode verwendet, bei dem die Codeelemente aus den Halbwellen bezw. Perioden eines sinusförmigen Wechselstromes mit den Kennzuständen kleiner und grosser Amplitudenwert bestehen. Einem Codeelement entspricht dabei einem bit. Werden z.B. 12 bit für das FBAS- und Tonsignal benötigt, so sind 12 Halbwellen erforderlich. Die Codierung kann synchron mit den Abgriffen bewerkstelligt werden, da sich die Länge der Codewörter sich nicht ändert. Wird dagegen ein Phasencode bezw. zusätzlich ein Phasencode vorgesehen, so ändert sich bei jeder Phasenänderung auch die Periodendauer, sodass bei einem periodischen Abgriff und bei gleichgerichteten Phasenänderungen die Signalabgriffe nicht mehr synchron mit dem Code sind. Zur Kompensation gibt es hier 2 Möglichkeiten - ausser einer Pufferspeicherung - einmal bei jeder Phasenänderung bis zur nächsten Phasenänderung die Nennfrequenz wieder herstellen, z.B. in der Fig.4 sie die Nennfrequenz f2 und erfolgt eine Phasenänderung T=f1 und haben die folgenden Codierungen dieselben Phasenänderungen, so werden die folgenden Codierungen mit der Nennfrequenz f2 codiert. Erst wenn sich die Phase f1 wieder ändert, erfolgt dann eine Phasenänderung in Bezug auf die Bezugsphase, d.h. beim Empfänger muss die Bezugsphase gespeichert werden. Diese kann z.B. in der Austastzeit vom Sender übertragen werden. Eine andere Möglichkeit Überlappungen zweier Abgriffe zu vermeiden, besteht darin, dass beim Sender mit jedem Codewort eine Messung zwischen Codewortende und dem vorhergehenden und dem folgenden Abgriff erfolgt. Ist die Gefahr einer Überlappung in voreilender oder nacheilender Richtung vorhanden, so werden Codewörter mit den kleinsten oder grössten Periodendauern zwischengeschaltet. In den Fig.29a und 29b sind solche dargestellt. Durch Zeilenspeicherung kann man dies umgehen.

In der Fig.19 hat ein Codeelement 6 verschiedene Stufen und 2 Stellen das Codewort, infolgedessen sind 6 hoch 2 Kombinationen möglich, also 36 Kombinationen. Mit 32 Kombinationen erhält man 5 bit. In der Fig.21b kann ein Codeelement ebenfalls 6 Stufen annehmen, sodass bei 5 Stellen 6 hoch 5 = 5184 Kombinationen möglich sind, also mindestens 12 bit. Bei 12 bit erhält man 4096 Kombinationen.

In der Fig.22 wird die PAM für den Ton im TSO-Glied erzeugt und jeweils z.B. halbzeilenweise an 6c gelegt. Die Anschlüsse 6c und 6d sind nicht erforderlich , wenn der Ton und die sonstigen Signale in die Austastzeit gelegt werden, sodass dann der Konzentator K1 diese Aufgaben übernimmt.

Mit Hilfe der Fig.21,22 und 23 sollte gezeigt werden. wie man z.B. den Codemultiplex auch beim Fernsehen anwenden kann. Die Übertragungsfrequenz kann natürlich wesentlich verkleinert werden, wenn man mehr Amplituden und/oder Phasenstufen vorsieht . Man kann auch zusätzlich mit verschiedenen Trägern, wie z.B. in der Patentanmeldung P 32 29 139.6 Fig.9 vorgesehen, oder mit verschiedenen Stromwegen kombinieren. So kann man z.B. in Fig.28 mit 8 KHz einen 64 Kbit Sprachkanal übertragen, und zwar mit einem Binärcode. 2 Stellen werden jeweils durch die beiden Halbwellen eines 8 KHz Wechselstromes markiert, 2 weitere Stellen durch die 2 Halbwellen eines Wechselstromes , der um 90 Grad phasenverschoben ist. Diese beiden Wechselströme werden summiert und als ein Wechstrom über den einen Stromweg übertragen. Dasselbe erfolgt über einen 2. Stromweg, sodass das Codewort 8-stellig und 2-stufig ist, sodass man 256 Kombinationen erhält. Auf der Empfangsseite wird nach der Auswertung der Halbwellen und natürlich Zwischenspeicherung eine Dekodierung vorgenommen. Die Codierung kann auch duobinär erfolgen.

Eine weitere Methode, insbesondere analoge Signale wie Sprache, Töne, das Luminanzsignal beim Fernsehen, die Farbsignale beim Fernsehen, Fernwirkwerte, frequenzmoduliert zu übertragen und zwar mit weniger Bandbreite, besteht darin mit Hilfe der Pulsdauermodulation PDM die Grösse der PAM-Impulse ind PDM Impulslängen umzuwandeln. Diese PDM-Impulse können dann in Wechselstromimpulse z.B. nach dem Verfahren der Fig.7 umgewandelt werden. Die Impulse werden dann durch die Halbwellen bezw. Perioden eines Wechselstromes gebildet, wobei die Periodendauern bezw. Halbperiodendauern der Halbwellen bezw. Perioden gleich der Länge der PDM-Impulse werden.

Das Spektrum der bisher verwendeten frequenzmodulierten Schwingung enthält oberhalb und unterhalb des Trägers eine grosse Anzahl von Seitenschwingungen, sodass ein sehr breites Band bei der Übertragung erforderlich ist. Die bnötigte Bandbreite ist dabei grösser als der doppelte Frequenzhub. Bei der erfindungsgemässen Schaltung können überwiegend digitale Schaltmittel verwendet werden, sodass eine preiswerte Herstellung möglich ist

Nachstehend wird nun die Methode an Hand von Zeichnungen näher erläutert. Zuerst werden bekannte Schaltungen nochmals erläutert,die u.a.bei der Erzeugung notwendig sind (Europäische Patentanmeldung 0 284 019). 2 Ausführungsbeispiele der Erfindung werden nachstehend beschrieben. Zuerst werden die Prinzipien der beiden Ausführungen zusammengefasst. Die Information wird einmal pulsamplitudenmoduliert und in der Folge mit Hilfe des Äquidestanzverfahrens In pulsdauern umgewandelt, oder aber die Information wird unmittelbar mit Hilfe des Sägezahnverfahrens In Pulsdauern codiert. Diese Pulsdauern werden dann in Verbindung mit den Pausen zwischen den Pulsdauern zu Rechteckimpulsen und in der Folge mit Hilfe von Filtern zu sinusförmigen Codierwechselströmen umgewandelt. Die Umformung der Pulsdauern und Pausen erfolgt mit Hilfe von Zählgliedern In Verbindung mit elektronischen Schaltern. Die Pulsdauer entspricht dann der Dauer einer Halbperiode bezw. Periode des Codierwechselstromes. Ist die Pulsdauer klein, ist die Frequenz der Halbwelle bezw. Periode beim Codierwechselstromes hoch, ist die Pulsdauer gross, so ist die Frequenz der Halbwelle bezw. Periode beim Codierwechselstrom klein. Auf der Empfangsseite erfolgt die Auswertung beispielsweise durch Abmessung der Halb- bezw. Periodendauern. Hier liegt also gleichzeitig eine Frequenz- und Phasenmodulation vor.

Bei der 2. Ausführungsform werden der Pulsdauerimpuls, in Fig 32 PD1,PD2 und die Pause zwischen den Pulsdauern (Fig 32,P) - die Pulsdauer und die Pause entspricht z.B. jeweils dem Abstand zwischen 2 Abgriffen, in Fig 30a mit tp bezeichnet.-einem elektronischen Relais zugeführt, in dem dann bipolare Rechteckimpulse erzeugt werden. Mit Hilfe von Filtern wird dann der frequenzmodulierte Codierwechselstrom erzeugt.

In der Fig.7 ist dargestellt wie mit Hilfe eines Zählgliedes Z in Verbindung mit der Frequenz der Fortschalte- bezw. Messimpulse , die im Oszillator Osc erzeugt werden, die Zeit eines Pulses bestimmt wird. Der jeweilige Ausgang des Zählgliedes markiert dann die Zeit. Dieser wird dann in Verbindung mit Gattern für die Steuerung eines elektronischen Relais ER vorgesehen. Dieses erzeugt dann bipolare Rechteckimpulse.

Die Funktion ist im Einzelnen folgende. Im Oszillator Osc werden die Fortschalte- bezw. Messimpulse für das Zählglied Z erzeugt. Diese gelangen über das Gatter G1 auf das Zählglied Z, solange

das Beginnzeichen an B vorhanden ist. Im Beispiel werden nur die Ausgänge Z1 und Z2 des Zählgliedes benötigt. Diese Ausgänge liegen an den Gattern G2 und G3. Soll die Halbperiode des Rechtimpulses J die Grösse der Summe der Messimpulse bis Z1 haben, wird vom Codierer Cod aus an g3 h-Potential gelegt, sodass beim Erreichen des Ausganges Z1 am Ausgang von G3 ein Potentialwechsel stattfindet, der das elektronische Relais ER veranlasst den Rechteckimpuls zu beenden. War dies ein positiver Impuls, so wird der nächste Impuls negativ. Das Zählglied wird dann in dieser Stellung wieder zurückgeschaltet. Am Ausgang z2 ist hierfür das Gatter G4 vorgesehen. Vom Codierer aus kann auch über fA die Oszillatorfrequenz vergrössert oder verkleinert werden, sodass man z.B. mit den jeweiligen Ausgängen verschiedene Zeiten markieren könnte. Vom Codierer Cod geht auch eine Verbindung A zu ER, mit der man verschiedene Impulsgrössen J steuern kann.

Die Rechteckimpulse werden über einen Tiefpass TP , den Übertrager Ü und Filter Fi als sinusförmiger Codierwechselstrom auf die Leitung gegeben. Die Halb- bezw. Periode des Codierwechselstromes ist dieselbe wie die des Rechteckimpulses. Das Prinzip der Umwandlung der Rechteckimpulse in einen sinusförmigen Wechselstrom ist in der Fig.3 dargestellt. Werden z.B. Rechteckimpulse mit der Frequenz 1 MHz mit einem Tiefpass 5,5 MHz bandbegrenzt, so erhält man, wie in der Fig.3c dargestellt ist,noch ziemlich steile Flanken. In der Fig.3b wurde ein Tiefpass von 3,5 MHz eingesetzt, man sieht, dass hier die Flankensteilheit schon merklich nachgelassen hat. In der Fig3 a ist ein Tiefpass von 1,5 MHz eingeschaltet, beim Empfänger hat man hier einen sinusähnlichen Wechselstrom. Die Periodendauern sind dabei die gleichen wie die der Rechteckimpulse, d.h. man kann die Periodendauern als Mass für die Frequenzen bezw. Phasen hernehmen. In der Fig.7 wurde dieses Prinzip bei der Umwandlung der Rechteckimpulse J in einen Codierwechselstrom mit Hilfe des Tiefpasses TP angewendet.

In der Fig.4 sind Rechteckimpulse verschiedener Periodendauern aufgezeichnet, und zwar durch die Frequenzen ausgedrückt f,f1 und f2. Diese Rechteckimpulse haben gegeneinander verschiedene Phasenverschiebungen bezw. verschiedene Frequenzen. Man sieht hieraus, dass man durch Änderung der Periodenauern Phasensprünge bezw. Frequenzsprünge hervorrufen kann, sodass man hierdurch auch eine Frequenzmodulation erhält. In der Fig.5 erfolgt solch ein Phasen- bezw. Frequenzsprung stufenweise. Damit wird erreicht, dass die Bandbreite klein wird. Wie aus der Fig 6 hervorgeht, erhält man bei Phasensprüngen von 5 Grad je 180 Grad bei 4 Phasensprungstufen eine Gesamtphasenverschiebung von 40 Grad.

In der Fig30a sind PAM-codierte Pulse von einem Signal Inf dargestellt. Diese werden mit Hilfe eines Äquidistanzverfahren in Pulsdauerimpulse , wie in der Fig 30b gezeigt ist, umgewandelt. Der Abstand der PAM-Impulse (Fig 30a tp) zueinander entspricht jeweils einer Pulsdauer PD und einer Pause P , wie in der Fig 30b dargestellt. Eine Pulsdauermodulation kann auch mit Hil fe des Sägezahnverfahrens durchgeführt werden. In den Fig.31 und 32 ist dieses Verfahren dargestellt. Die Pulsdauern sind Rechteckpulse PD1,PD2 ,...Weiterhin sind bekannt die symmetrische PDM und die bipolare PDM. (siehe auch Buch "Modulationsverfahren" von Stadler 1983).

In der Fig.35 ist ein Ausführungsbeispiel gemäss der Erfindung dargestellt. Im Pulsdauermodulator PDM werden die Pulse z.B. nach Fig 30b oder 32 erzeugt, und über G5 an das Gatter G1 geführt.Am anderen Eingang des Gatters G1 liegen die Messimpulse Jm, z.B. 100KHz Frequenz. Solange an G1 ein PD-Puls liegt, werden die Messimpulse Jm am Ausgang wirksam. Über das Potentialumkehrgatter G2 gelangen die Messimpulse an das Zählglied Z, das mit diesen Impulsen gesteuert wird. Die Zahl der Ausgänge am Zählglied entspricht z.B. dem Abstand zwischen 2 PAM-Pulsen, in Fig 30a tp. Die Abgriffsfrequenz sei 10 Khz, dann hätte das Zählglied 100.000 Ausgänge. Der Frequenzhub wird durch den grössten und kleinsten Amplitudenwert der Information Inf bestimmt, in Fig 30a mit gw und kw bezeichnet. Die Ausgänge A des Zählgliedes Z führen zu Gattern G3 und die Ausgänge der Gatter zu Gattern G4. Jeweils am anderen Eingang des Gatters G4 liegt der jeweilige PD-Impuls, der das Gatter G4 sperrt. Erst wenn der PD-Impuls nicht mehr da ist, kann auch das Ausgangspotential über G3 an G4 wirksam werden. ER erhält nun über G4 ein Potentialwechselkennzeichen für den nächsten Rechteckimpuls. Der Beginn des Rechteckimpulses wird durch den jeweiligen PD-Puls markiert. Der nächste Rechteckimpuls wird durch die Pause P (Fig 30b P) bestimmt. Von ER wird über P ein Potential an Gatter 5 gelegt, damit am Gatter G1 die Messimpulse Jm wieder durchlässig werden. Das Zählglied Z wird nun bis zum Ausgang Gatter G6 geschaltet. Wenn der nächste PD-Puls wieder kommt wird G6 wirksam und über R wird das Zählglied wieder in die Ausgangsstellung geschaltet. Am Ausgang von ER sind dann Rechteckimpulse RJ der Grösse der Halbperioden wie die der PD- pulse und der Pausen P. Im Filter Fi werden die Rechteckimpulse zu sinusförmigen Halbwellen fmo, damit ist die Information frequenzmoduliert. Die Halbperioden der Nutzsignalmodulationsfrequenzen bewegen sich dann zwischen den Halbperiodendauern am Zählglied mit kw und gw gekennzeichnet. In Fig. 33 ist z.B. kw = 15 KHz, die Mittenfrequenz 10 KHz und in

Fig.·34 gw = 75 KHz. Im Beispiel können sich die Pulsdauern um die Hälfte ändern. dies ist eine Dimensionierungsache der Pulsdauermodulationsschaltungen. Die Halbwellen der Pausen haben in der Fig. 33 eine kleinste Frequenz von 7,5 KHz und in Fig.34 eine grösste Frequenz von 15 KHz. Die Amplituden der Halbwellen bleiben immer gleich. Die Auswertung auf der Empfangsseite erfolgt durch Abmessung der Halbperiodendauern. Eine Synchronisierung ist nicht erforderlich, da die Nulldurchgänge einer Periode bei einer Codierung mit Hilfe einer PAM zugleich die Abgriffe codieren, es müssen also lediglich die positiven Halbwellen in PAM-Pulse umgewandelt werden. Die PAM-Pulse sind dann auf der Empfangsseite um eine periode nacheilend.

Die Redundanz der Pausen in der Fig.35 kann vermieden werden, wenn man z.B. die PAM-Pulse speichert und nach jeder PD-Codierung den nächsten PAM-Puls abruft. Beim Empfänger Ist allerdings dann eine Synchronisierung erforderlich. Bei Verwendung der PAM auf der Sendeseite müsste die Abgriffsfrequenz von Zeit zu Zeit synchronisiert werden. In Fig. 36 ist die Prinzipschaltung einer solchen Schaltung auf der Sendeseite dargestellt. Die PAM-Pulse werden Im Speicher Sp gespeichert Von ER kommt über AR der Abruf des nächsten Impulses. Vorbereitend war schon der nächste Impuls als PDM-Impuls im Speicher Sp1 gespeichert. Damit wird nun über das Steuerorgan St das Zählglied Z gesteuert und auf einen entsprechenden Ausgang eingestellt. Von ER wurde auch über R das Zählglied wieder in die Ausgangsstellung gebracht. Am Steuerorgan liegen auch die Steuerimpulse Jm. Mit dem Abruf des PDM-Impulses wird auch vom Speicher Sp ein PAM-Impuls zum Pulsdauermodulator gegeben und in diesem als PDM-Impuls solange gespeichert, bis der Sp1 Speicher wieder frei ist. Zweckmässig wird man 2 Sp1 Speicher vorsehen, die dann abwechselnd an das Steuergerät nach jedem Abruf von ER gelegt werden. Am Ende des PDM-Impulses wird über das Zählglied Z,G1,G2 ein Impuls-Endekriterium an ER gegeben. Der von ER erzeugte Rechteckimpuls PD wird auf den nächsten umgepolt, über R das Zählglied zurückgeschaltet und über AR der Abruf des näch-

In der Fig.39 sind 4 Kanäle dargestellt mit einer Halbwellencodierung mit den Kennzuständen grosser und kleiner Amplitudenwert.Für alle 4 Kanäle ist die Frequenz die gleiche. Diese 4 Kanäle werden für die Codierung der Farbfernsehsignale vorgesehen. 8 bit sind für das Y-Signal (Luminanzsignal) und zwar je 4 bit beim Kanal a und b. je 2 bit in den Kanälen a und b sind für Ton und sonstige Signale T + S vorgesehen. Der Kanal c ist für die Codierung des rot-Signales und der Kanal d für die Codierung des blau-Signals mit je 6 bit vorhanden. Je 2 Kanäle werden dann entsprechend der Fig. 11 Vektor I,(k1,k2) mit den Codierungen I,(II),IV, (III) zusammengefasst, sodass ein Summenwechselstrom entsprechend der Fig.9 zustandekommt. Die Phasenlage der beiden Summenwechselströme wird dann auf 0 Grad und 90 Grad festgelegt. Diese beide Summenwechselströme kann man nun auf der Basis der Quadraturamplitudenmodulation übertragen, sodass für die Übertragung aller Farbfernseh- und sonstigen Signale ein schmales Band benötigt wird. Als doppelte QAM übertragen, d.h. Kanal a + b quadraturamplitudenmoduliert und die Kanäle c + d quadraturamplitudenmoduliert, wobei die Kanäle zueinander 0°,90°,90° und 180° Phasenlage aufweisen und deren Summenwechselströme 45° und 135° Phasenlage haben, und dass die beiden Summenwechselströme wieder quadraturamplitudenmoduliert werden, ist die Auswertung schwieriger, wie auch aus der Fig.11 ersichtlich ist (bei einmaliger QAM entstehen die Vektoren I,II und III).

Man kann die 4 Kanäle bezw. ihre binäre Werte auch codemultiplex übertragen. In der Fig.40 sind die Binärwerte der 4 Kanäle nochmals dargestellt. Entsprechend der Fig.41 sollen jeweils 2 Reihen der Fig.40 zu 8 bit zusammengefasst werden. In der Fig.39 sei 6 MHz die Frequenz der Wechselströme, für die Codierung sind dann 18 MHz erforderlich. Verwendet man in der Fig.41 eine duobinäre Codierung entsprechend der Fig. 62 mit den Halbwellen als Codeelemente, so würde man zwar gegenüber der Fig.39 an Bandbreite etwas gewinnen, aber die Frequenz wäre 3mal so hoch. Fasst man die Reihen 1,2,3 und 4,5,6 , also 12 bit jeweils zusammen bei diesem duobinären Code, so ist für eine Reihe 1,2,3 ein 3-stufiges Codewort mit 8 Stellen erforderlich. 8 Stellen bedeuten 4 Perioden, Es wären also eine Frequenz von 2x24 MHz erforderlich, al so auch für diesen Zweck zu hoch. In der Fig.45 ist ein 4-stufiges Codeelement dargestellt. bei 4 Stellen ergibt dies 256 Möglichkeiten. Eine Codierung nach Fig.41 ergäbe eine Frequenzreduzierung auf 36 MHz. In der Fig.63 ist ein 6 stufiges Codeelement dargestellt. Um 3 Reihen der Fig.40 seriell zu codieren , also 12 bit, wären hier 5 Stellen erforderlich. Es wären also noch 30 MHz erforderlich. Ausser den 3 Amplitudenstufen sind noch zwei Phasenstufen bezw. Periodendauern vorgesehen. In der Fig.46 sind 3 Amplituden und 3 Phasenstufen dargestellt.Werden aus der Anordnung der Fig.40 2 Reihen mit je 12 bit gebildet, sind für jede Reihe 3 Stellen erforderlich, für beide Reihen also 6 Stellen, d.h. es ist eine Frequenz von 18 MHz notwendig.

In der Fig.43 sind die Farbfernsehsignale an-

derst angeordnet. 8 bit für einen Y-Abgriff (Luminanz,Bildpunkt B) sind seriell zu je 4 bit, die Farben rot oder blau seriell je 3 bit in den Reihen III + IV. Das jeweils 4.bit in den Reihen 3 und 4 ist für Ton- und andere Zwecke vorgesehen. Die Farbe rot oder blau kommt jeweils bei jedem 2. Y-Signal, d.h. diese wechseln sich laufend ab. Werden die senkrechten Reihen 1/2 und 3/4 , wie in der Fig.44 dargestellt, zusammengefasst, so ergeben sich bei einer Codierung günstigere Verhältniss. Bei 4 Stufen sind 3 Stellen erforderlich, es ist dann eine Frequenz von 18 MHz erforderlich. Werden die Reihen 1/2 und 3/4 parallel angeordnet, also 16 bit, so sind bei einer Codierung nach Fig.46 4 Stellen erforderlich, also 12 MHz Frequenz. Die doppelte QAM der Fig.39 kann, um noch mehr Sicherheit bei der Übertragung zu haben, frequenzmoduliert übertragen werden. Der Summenwechselstrom hat nur kleine Frequenzänderungen, sodass, wie aus der Fig.64 hervorgeht, die frequenzmodulierte Schwingung doch schmalbandig übertragen werden kann. Aus dieser Fig. geht hervor, dass die Halbperiodendauer T/2 bei einer Frequenzerhöhung sehr klein wird, dass also die Frequenz stark zunimmt. Bei einer Modulationsfrequenz Mf und einer Amplitude u ist die Halbperiodendauer T/2, bei doppelter Amplitude 2u ist die Halbperiodendauer kleiner, während bei zusätzlich doppelter Frequenz M2f sich die Halbperiodendauer wesentlich verkleinert.

In der Fig.47 ist eine Übersicht über einen Fernsehsender dar gestellt, bei der die in den Fig. 40,41,43 und 44 erläuterten Codes verwendet werden. Vom Multiplexer (nicht einge zeichnet) kommen die analog abgegriffenen Signale in den Analogspeicher ASp und von dort werden die Probeentnahmen an einen oder mehrere Analog/Digitalwandler weitergegeben. Die digitalisierten Signale werden dann im Digitalspeicher DSp gespeichert und in der Folge dem Ordner zugeführt. In diesem werden sie entsprechend den Fig.40,41,43 oder 44 geordnet. So geordnet werden sie dem Codierer zugeführt. Entsprechend dem vorbestimmten Code z.B. nach Fig.45 oder 46 oder 62 oder 63 codiert und dem Modulator MO zugeführt-Vom Oszillator wird der Sendewechselstrom dem Modulator zugeführt und der modulierte Sendewechselstrom über nicht eingezeichnete Verstärkerstufenund dem Endverstärker zur Antenne gegeben. Eine Übersicht vom Empfänger für die Auswertung der codierten Signale ist in der Fig. 48 dargestellt. Der Sebdewechselstrom kommt über die Empfangsantenne E in die Stufen Abstimmkrois/Verstärker, Mischstufe/Oszillaotr Mi/Osc , über den Zwischenfrequenzverstärker ZF zur Demodulationsstufe - der Eingang ist wie ein Überlagerungsempfänger beim Rundfunkempfang geschaltet- , am Ausgang des Demodulators ist der Codewechselstrom vorhanden. Dieser wird in den Decodierer geschaltet Die im Sendemultiplexer abgegriffenen Signale werden hier wieder erhalten,wie das Y, r-y, b-y, Ton und sonstigen Signale S und den verschiedenen Schaltungen zugeführt.

In den Fig. 50 und 51 sind analoge Codierungen der Farbfernsehsignale dargestellt. In der Fig.50 ist ein Wechselstrom gleicher Frequenz als Codewechselstrom vorgesehen. Die Amplituden der Halbwellen sind die Codeelemente. Die Abgriffsfolge ist y,r,y,bl,y,T + S usw. Die Übertragung dieser analog codierten Signale erfolgt auf der Basis der Frequenzmodulation, sodass man ein schmales Band - nur eine Frequenz Fig. 64 -und auch eine Übertragungssicherheit erhält.

In der Fig.51 wird ebenfalls ein Analogcode vorgesehen. Es ist eine Phasencodierung. Der Analogcode ist durch verschieden grosse Halbperiodendauern gegeben. Die Amplituden der Halbwellen haben dabei immer dieselbe Grösse, es ist eine Art Frequenz- und Phasenmodulation. Die einzelnen Signale sind wieder seriell angeordnet , im Beispiel y,r,y,bl,y,T + S. Die Übertragung erfolgt bei einer Abgriffsfrequenz des Y__-Signales mit 6 MHz mit 6 MHz. Erfolgt ein Multiplexabgriff aller Signale, also auch des r,bl und T + S Signale , so ist eine Abgriffsfrequenz von 12 MHz erforderlich.

In der Fig.52 ist eine Codierung entsprechend der Fig.51 vorgesehen, lediglich die Ton und sonstigen Signale T + S werden durch einen überlagerten Amplitudencode codiert. Es ist ein Binärcode mit einer grossen und einer kleinen Amplitude. Die Werte des Y und der r + bl-Signale sind durch die Halbperiodendauern festgelegt. Synchron mit dem PDM-Impuls wird dann z.B. an das ER-Relais der Fig.36 der jeweilige Amplitudenwert gegeben in dem dann ein Rechteckimpuls mit kleiner oder grosser Spannung erzeugt wird. Die Amplitudencodeelemente können z.B. mehreren Kanälen, wie Ton Stereo usw. zugeordnet sein. In der Fig.55 sind die 4 Halbwellencodeelemente 4 verschiedenen Kanälen zugeordnet.

Eine Auswertung der PDM, PPM oder PFM-Impulse mit den Halbperiodendauern codiert, istaus der Fig. 59 ersichtlich. Diese erfolgt wieder mit Hilfe einer Sägezahnspannung. Beim Beginn einer Halbwelle, also beim Nulldurchgang wird der Erzeuger der Sägezahnspannung eingeschaltet , nach der Halbwelle beim nächsten Nulldurchgang wird z.B. mittels eines Feldeffekttransistors die Sägezahnspannung kurzzeitig an einen Kondensator geschaltet und in diesem gespeichert. Die Halbperiodendauer T/2 ist dann gleich dem Spannungswert T/2 oder analog der Grösse des Spannungswertes. Die Halbperiodendauer von 1 entspricht dem Spannungswert u1, die von 2 dem von u2, usw. Wurde auf der Sendeseite Sprache mit 8 KHz pulsamplitudenmoduliert, so muss auf der Emp-

fangsseite mit derselben Frequenz die Spannung u1.u2.u3 jeweils abgegriffen werden und zum Sprachwechselstrom umgeformt werden. Bei einem zeitmultiplexen Abgriff mehrerer Kanäle, müssen die gespeicherten Werte u1,u2,u3,... mit derselben Frequenz des zeitmultiplexen Abgriffes wieder verteilt werden. Die Herstellung der ursprünglichen Information kann z.B. in der Weise erfolgen, indem man den ausgewerteten Code u1,u2,.. nach der Kanalzuteilung treppenförmig ausbildet und dieses Treppensignal über einen Tiefpass führt. Solche Umfor mungen sind bekannt und es wird daher nicht näher darauf eingegangen.

Auf dieselbe Weise wie in Fig.59 die PDM-Impulse können auch PPM-Impulse decodiert werden. In der Fig.60 ist dies dargestellt. Der Abstand T 2 der pulse wird mit der Sägezahnme thode wieder in PAM-Pulse umgeformt und gespeichert. Der Abstand T 2 entspricht dann der Spannung u1 usw.

Bei der Übertragung von Fernsehsignalen nach dem Prinzip der Fig.36 und 38 müssen die ausgewerteten Signale auf der Empfangsseite synchron verteilt werden. In der Austastzeit müssen synchronisierimpulse gesendet werden, damit entsprechend der Sendeseite die Abtastfrequenz auf der Empfangsseite die Verteilfrequenz festgelegt werden kann. Die Summe der vorkommenden grössten Halbperiodendauern je Zeile darf die Zeit von 54 us nicht überschreiten. dies ist die Zeit die für eine Zeile bei einem Bildformat 4:3 vorgesehen ist. Im Sender müssen infolgedessen die Halbperiodendauern mit abgemessen werden u.U. muss in den Zeilencode noch ein Füllcode, der z.B. die kleinsten oder grössten Periodendauern in bestimmter Folge beinhaltet. Man kann natürlich auch andere Füllcodes vorsehen. Ausserdem ist zusätzlich die Austastzeit als Füllcode noch vorzusehen. In der Fig.61 sind die kleinsten und grössten Halbperiodendauern k und g dargestellt. Solche können z.B. abwechselnd gesendet werden. Auf dieser Basis können auch mehrere Kanäle über einen Übertragungsweg zusammengefasst werden. In der Fig.56 ist ein solches Beispiel dargestellt. Mit dem Multiplexer Mu werden die Kanäle 1 bis n pulsamplitudenmässig zusammengefasst, was ja bekannt ist. Diese PAM-Proben werden im Speicher Sp gespeichert, vom PDM abgerufen und,wie bereits beschrieben, über ein Steuergerät St, an das die Steuerimpulse Jm angeschlossen sind, dem Zählglied zugeführt. Die übrigen Schaltvorgänge sind dieselben wie z.B. in der Fig.36 beschrieben. Nach dem Pulsdauermodulator PDM können die Impulse auch direkt entsprechend der Fig.38 weiter verarbeitet werden. Auf der Empfangsseite muss natürlich entsprechend der Abgriffsfrequenz des Multiplexers synchronisiert und verteilt werden.

In der Fig.57 ist eine andere Möglichkeit der Mehrfachausnutzung eines Stromweges aufgezeigt. Um die Codewechselströme frequenzmässig trennen zu können, werden solche Steuerimpulse verwendet, dass die Frequenzbereiche der Codewechselströme einen solchen Abstand haben, dass eine einwandfreie Auswertung möglich ist, z.B. mittels Filter eine Trennung in der Empfangsstelle. In der Fig.57 ist Z1 der eine Umsetzer mit den Steuerimpulsen Jm1 und Z2 der andere Umsetzer bezw. Zählglied mit den Steuerimpulsen Jm2. In der Fig.58 ist die Frequenzlage der beiden Kanäle dargestellt. T/2I und T/2II sind die kleinsten Frequenzen der beiden Kanäle . Durch den Winkelhub f2 kommt man näher an den Frequenzbereich vom Kanal T/2I. Im Beispiel ist noch ein Abstand von Ab vorhanden. Dieser kann so gewählt werden,dass preislich günstige Filter eingesetzt werden können.

Nachstehend werden noch einige Codes dargestellt, mit den man mit einer Frequenz Daten, im Beispiel Fernsehsignale codieren und übertragen kann. In der Fig.53 ist ein Binärcode dargestellt, bei dem als Codeelemente die Amplituden von Halbwellen mit den Kennzuständen grosser und kleiner Amplitudenwert vorgesehen werden. Mit einer Halbwelle kann dan ein bit codiert werden. Für das Y-Signal sind 8 bit, für das rot und Blausignal je 6 bit und für den Ton (digitalisiert) und sonstige Signale sind 2 bit vorgesehen. Rot und blau werden abwechselnd , wie z.B. in der Fig.51 dargestellt, codiert. Bei 6 Meg Abgriffen für das Y-Signal wäre hier ein Codierwechselstrom mit 48 MHz erforderlich. In der Fig.54 ist eine duobinäre Codierung hierfür vorgesehen. Der Codierwechselstrom hat dann eine Frequenz von 27 MHz. Man kann diese Codierwechselströme wieder frequenzmoduliert übertragen, das Frequenzband wird dabei auch nicht zu breit, wie aus der Fig. 64 hervorgeht. Die Übertragungssicherheit wird dabei noch grösser. In der Fig.66 ist eine Möglichkeit aufgezeichnet, wie man ohne Modulatoren schmalbandig eine Nachricht digital übertragen kann. Jedem Codeelement wird eine Vielzahl von Perioden eines Wechselstromes einer Frequenz zugeordnet, die durch die Zeit Og bestimmt werden, also einer vorbestimmten Zahl von Perioden. Angenommen wird die Codierung erfolgt binär. Bei jedem Zustandswechsel, also 1 nach 0 oder 0 nach 1 erfolgt der Übergang kontinuierlich , in der Fig.66 mit Ü bezeichnet. Die Amplituden für die Null haben die Grösse Ak und die für die 1 Ag. Kommen gleiche Werte hintereinander,so wird die Amplitudengrösse nicht geändert, bei 5 gleichen Werten würde man 5mal eine Periodenzahl von Og mit derselben Amplitude erhalten. Der Übergang zu einem anderen Kennzustand wird z.B. zur folgenden Kennzustand gerechnet, also z.B. Ü + O = Og. In der Fig.65 ist aufgezeichnet wie man seriell die Fernsehsignale digital anordnen kann.

In den Fig.53,54 und 66 sind die Frequenzbänder für die Übertragung der Fernsehsignale sehr schmal. U.u. könnte man Kanäle zwischen die einzelnen Fernsehkanäle unterbringen. In der Fig.42 ist hierfür der Träger BTz vorgesehen . Bei der Codierung nach der Fig.66 ist der Träger zugleich das Modulationssignal. Bei der Modulation des BAS-Signals mit dem Zwischenfrequenzträger 38.9 MHz wird ausser dem Filter für die Erzeugung des Restseitenband ein Saugkreis bezw. Reihenresonanzkreis in eine solche Frequenzlage gebracht, dass eine Kurve RR wie in der Fig.42 dargestellt, zustandekommt. Solch ein Reihenresonanzkreis ist leicht zu realisieren. Die Nyquistflanke dürfte durch diese Massnahme kaum beeinflusst werden.

Ansprüche

1. Verfahren für die digitale und/oder analoge Codierung von Information eines, zweier oder mehrerer Kanäle und/oder Frequenz oder Bandbreitenreduzierung und/oder Erhöhung der Übertragungssicherheit, dadurch gekennzeichnet, dass die Übertragung von Information eines, zweier oder einer Vielzahl von Kanälen mit weniger Bandbreite als der Einzelkanal bezw. die Summe der Bandbreiten zweier bezw. einer Vielzahl von Kanälen ausmächt, in der Weise erfolgt, indem die synchron bezw. quasisynchron angeordneten Codeelemente der zu übertragenen Kanäle parallel geordnet werden (Fig.20, S1,S2,..) und so zusammen zu einem Codewort vereinigt werden und/oder dass die zu codierende digitale oder analoge Information ggf. unter Zwischenschaltung von Zwischenstufen (z.B.PAM) in PDM-Impulse umgewandelt werden, dass weiterhin Mittel vorgesehen werden, die die Werte der PDM-Impulse in die Halbperiodenbezw. Periodendauern von Halbwellen oder Perioden eines sinusförmigen oder sinusähnlichen Wechselstromes umwandeln (Fig.35, ER,Fig. 36, ER, Fig.38 ER)

2. Verfahren zur Erzeugung einer Frequenzmodulation, dadurch gekennzeichnet, dass Mittel vorgesehen sind, die eine Information bezw. Signal (Fig.30a,Inf) in Pulsdauern umwandeln (Fig.30b,32), dass weiterhin Schaltmittel für die Abmessung der Pulsdauern, insbesondere Zählschaltmittel (Fig.35,Z) vorgesehen sind, die zugleich eine Markierung der Pulsdauern vornehmen(z.B. Fig.35,,Z,A), die Markierstromkreise sind dabei so in Verbindung mit Pulsdauerimpulsen über Gatter mit einem elektronischen Schaltmittel (Fig.35,ER) verbunden, dass der Anfang und das Ende des jeweiligen Pulsdauerimpulses ein periodisches Signal, insbesondere Rechteckimpuls, codieren, weiterhin sind solche Siebmittel vorgesehen, dass an die Leitung nur sinusähnliche bezw. sinusförmige Wechselströme oder/und oberwellen davon gelängen (Fig.35,fmo).

3. Verfahren zur Erzeugung einer Frequenzmodulation, dadurch gekennzeichnet, dass Mittel vorgesehen werden, die eine Information bezw. ein Signal in Pulsdauern umwandeln und dass weiterhin Schaltmittel vorgesehen werden, die die Dauerimpulse in eine ununterbrochene Folge (Pd,Pd,Pd,..) oder die die Pulsdauerimpulse und die dazugehörigen Pausen (Fig.32, PD1,P, PD2) in insbesondere Rechteckimpulse umwandeln (Fig.36,38) und dass in der Folge solche Siebmittel vorgesehen werden, die diese in sinusförmige oder sinusähnliche Halbwellen bezw. Perioden zu einem Codierwechselstrom umwandeln.

4. Verfahren nach den Ansprüchen 1 bis 3, dadurch gekennzeichnet, dass die Pulsdauerimpulse und Pausen bezw. bei Speicherung Pulsdauerimpulse in einer ununterbrochenen Folge elektronische Schaltmittel unmittelbar so steuern (ERFig.36,38), dass die jeweilige Pulsdauer bezw. Pulsdauerpause in eine Periodendauer bezw. Halbperiodendauer von unipolaren oder bipolaren Rechteckimpulsen umgewandelt wird und dass Siebmittel vorgesehen werden, die aus den Rechteckimpulsen sinusähnliche Halbwellen bezw. Perioden in einer ununterbrochenen Folge von positiven und negativen Halbwellen machen.

5. Verfahren zur Auswertung von Abständen z.B. zwischen Pulsen oder von Halb-oder Periodendauern, dadurch gekennzeichnet, dadurch gekennzeichnet, dass beim Anfang der Abstandsmarkierung (Fig.60,1) bezw. beim Nulldurchgang der Halbperiode Mittel zur Erzeugung einer Sägezahnspannung angelassen werden und dass am Ende der Abstandsmarkierung bezw. beim 2. Nulldurchgang der Halbperiode (Fig.59) Mittel an die Sägezahnspannung geschaltet werden die eine Abmessung derselben oder dass Mittel vorgesehen werten (FET) die diese Spannung insbesondere in einem Kondensator speichern.

6. Verfahren nach den Ansprüchen 1 bis 5, dadurch gekennzeichnet, dass eine Mehrfachausnützung von Stromwegen in der Weise erfolgt, indem mehrere Informationskanäle zeitmultiplex zusammengefasst werden (Fig.56) oder indem die Steuerimpulse für die Zählglieder eine solche Frequenz erhalten (Fig.57,Jm1,Jm2) dass ihre Codierwechselströme bei der Übertragung über einen Stromweg keine Überlappung erhalten.

7. Verfahren nach Anspruch 1, dadurch gekennzeichnet, dass für die Codierung ein mehrstufiger Amplitudencode (binär,duobinär usw.) und/oder ein oder mehrstufiger Phasencode und/oder ein analoger Amplituden und/oder Phasencode vorgesehen wird, der insbesondere für die Mehrfachausnutzung oder Verkleinerung der Frequenz beim Te-

lex (Fig.18.19.20) beim Fernsehen (Fig.21) bei Teletex. Datenübertragung (Fig.24) bei der digitalen Sprachübertragung (Fig.28) vorgesehen wird.

8. Verfahren für das Farbfernsehen, dadurch gekennzeichnet, dass auf der Sendeseite alle Signale codemultiplex zusammengefasst werden , wobei die Farb- Ton- und sonstigen Signale codemultiplex mehreren Y-Signalen bedarfsweise zugeordnet werden können und dass die Empfangsseite wie ein Überlagerungsempfänger (Superheterodyn) ausgebildet ist wobei hinter dem Demodulator (Fig.23.DM) der Decodierer angeordnet ist mit dem zeitgerecht die decodierten Signale verteilt werden.

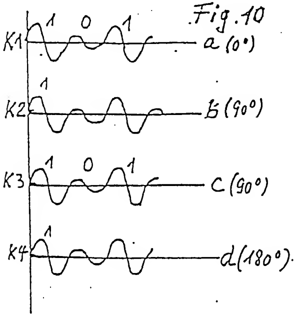
9. Verfahren für die Codierung der Farbfernsehsignale . dadurch gekennzeichnet, dass seriell das y-Signal. rot-Signal y-Signal, Blausignal, Y-Signal,Ton + sonstigen Signaleabgegriffen werden In einer ununterbrochenen Reihenfolge, dass die PAM-Werte auf die Halbperioden- bezw. Periodendauer von Halbwellen bezw. Perioden eines Wechselstromes übertragen werden und zwar bei Amplitudengleichheit oder dass nur die Reihenfolge Y.r.Y.bl vorgesehen wird und die Ton- und sonstigen Signale durch einen binären bezw. duobinären Amplitudencode (Fig.55) in der Weise codiert wird, indem jeder Halbwelle oder Periode ein dem Code entsprechender Amplitudenwert zugeordnet wird , wobei die 4 Amplitudenwerte (Fig. 52) codemultiplex verschiedenen Kanälenzugeordnet werden können.

10. Verfahren für die Codierung der Farbfernsehsignale , dadurch gekennzeichnet, dass die Fernsehsignale nur mit einer Frequenz (Fig.53.54.66) in der Weise codiert werden, indem die seriell angeordneten Codeelemente , die durch die Amplituden der Halbwellen bezw. Perioden mit den Kennwerten grosser oder kleiner Amplitudenwert oder kleiner,mittlerer und grosser Amplitudenwert gebildet werden für alle Signale vorgesehen werden oder dass der Code aus einer Vielzahl von Perioden gebildet wird mit 2 oder 3 Kenngrössen und einem kontinuierlichen Übergang zwischen den Grössen (Fig.66,Ü), wobei bedarfsweise die ser Code für die Unterbringung eines Kanals in der Lücke zwischen den herkömlichen Kanälen vorgesehen ist.Fig.42).

11. Verfahren nach den Ansprüchen 1,7,9 und 10, dadurch gekennzeichnet, dass die Auswertung auf der Empfangsseite bis zum Decoder wie bei einem Überlagerungsempfänger erfolgt.

12. Verfahren nach den Ansprüchen 1,7,8 bis 11, dadurch gekennzeichnet. dass eine Übertragung der Fernsehsignale auf der Basis der doppeltem QAM erfolgt. wobei das y-Signal auf 2 Kanäle mit je 4 bit verteilt wird und diesen Kanälen zusätzlich je 2 bit für Ton- und sonstigen Zwecke zugeordnet wird. die Codeelemente sind die Halbwellen eines Wechselstromes mit den Kennzuständen grosser und kleiner oder grosser,mittlerer und kleiner Amplitudenwert, die Übertragung erfolgt bedarfsweise auf der Basis der Frequenzmodulation.

Fig. 10



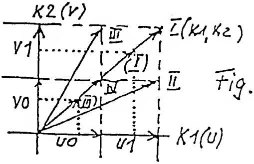


Fig. 11

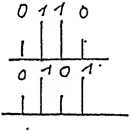


Fig. 13

Fig. 1:

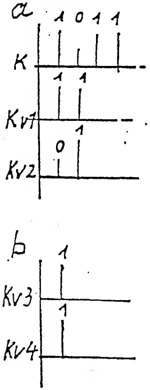


Fig. 31a

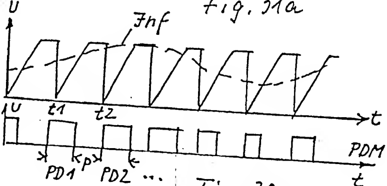


Fig. 32a

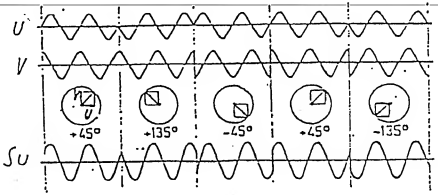


Fig. 2

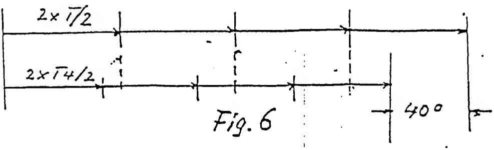


Fig. 6

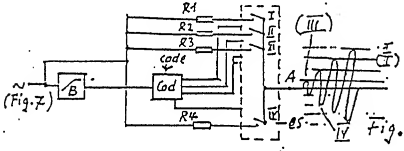


Fig. 9

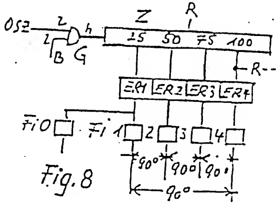


Fig. 8

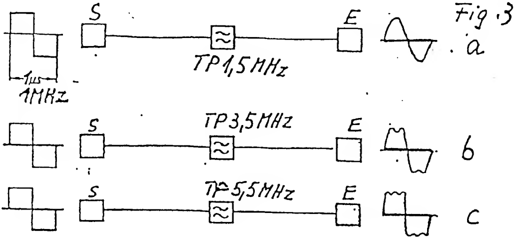
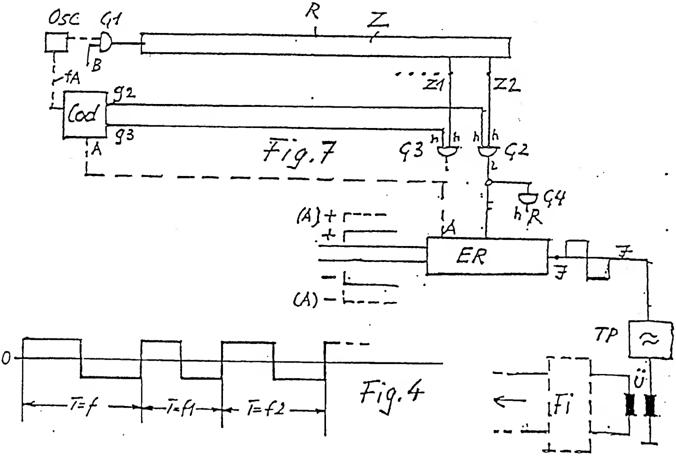
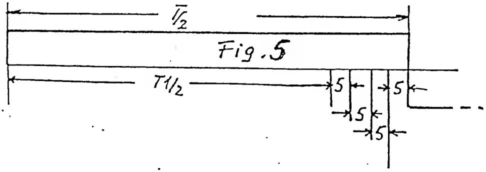


Fig.3
S
E
TP 1,5 MHz
1µs
1MHz
a
TP 3,5 MHz
b
TP 5,5 MHz
c


Osc
G1
R
Z
B
fA
Cod
g2
g3
A
Z1
Z2
G3
G2
G4
h
Fig.7
(A) +
+
−
(A) −
ER
F
TP
Ü
Fi
0
T=f
T=f1
T=f2
Fig.4


T/2
Fig.5
T1/2
5

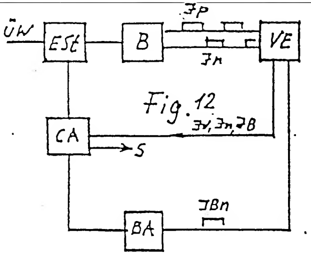


Fig. 12

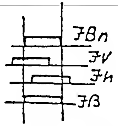


Fig. 12a

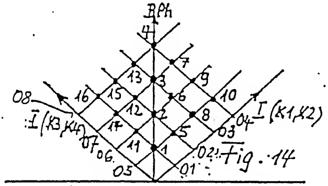


Fig. 14

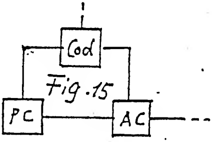


Fig. 15

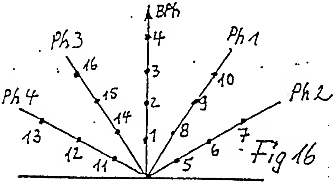


Fig 16

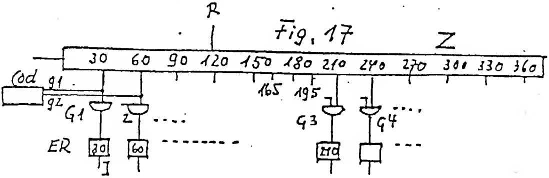


Fig. 17

Fig.18

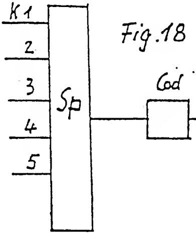


Fig.19

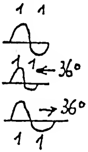
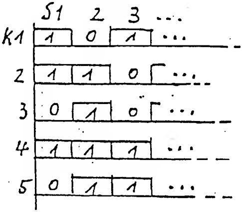


Fig.20



$S1 = 1-1-0-1-0$

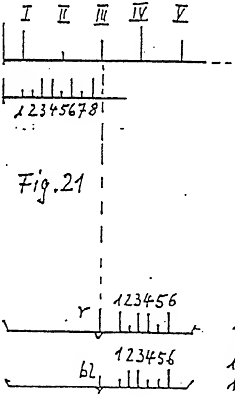
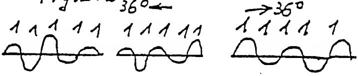


Fig.21

Fig.21b



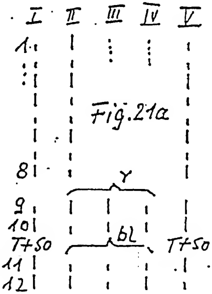


Fig.21a

Fig. 22

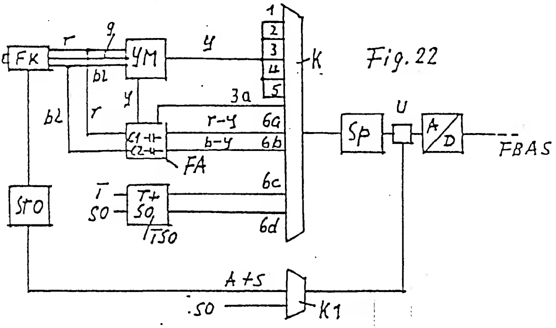


Fig. 23

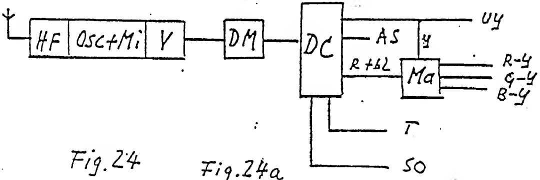


Fig. 24

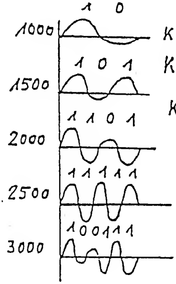


Fig. 24a

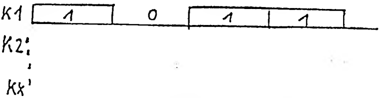


Fig. 29a

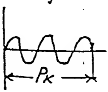
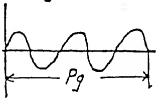


Fig. 29b



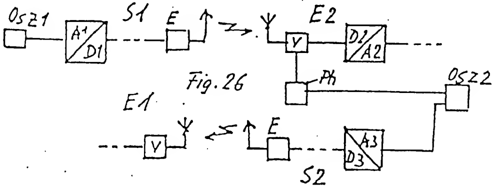


Fig.26

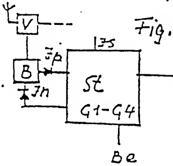
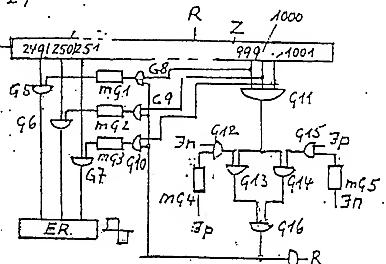



Fig. 27

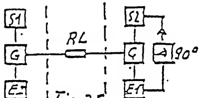


Fig.25

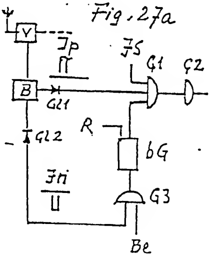
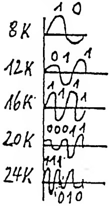


Fig.27a

Fig.28



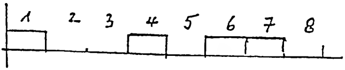


Fig. 28a

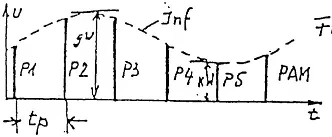


Fig. 30a

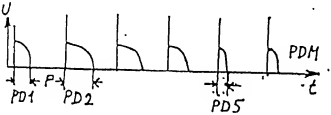


Fig. 30b

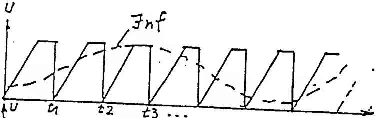


Fig. 31

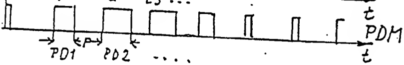


Fig. 32

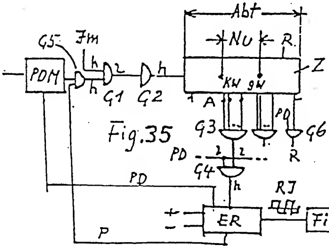


Fig. 35

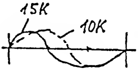


Fig. 33

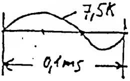


Fig. 34

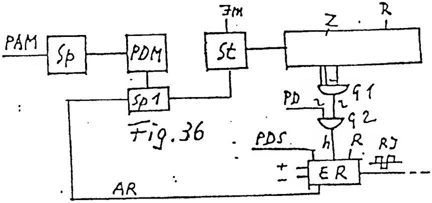


Fig. 36

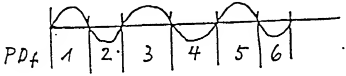


Fig. 37

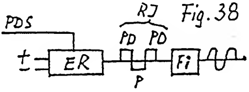


Fig. 38

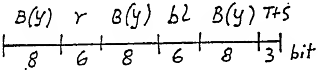


Fig. 65

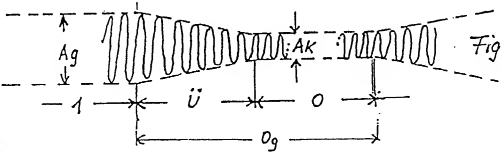


Fig. 66

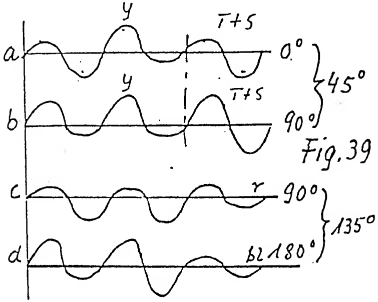
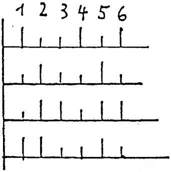


Fig.39

Fig. 40



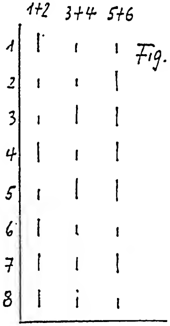


Fig. 41

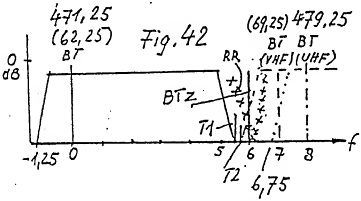


Fig. 42

Fig. 43

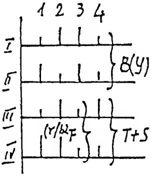


Fig. 44

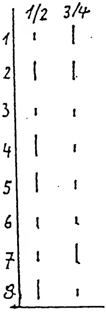


Fig. 45

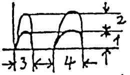


Fig. 46

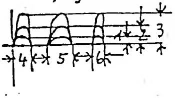
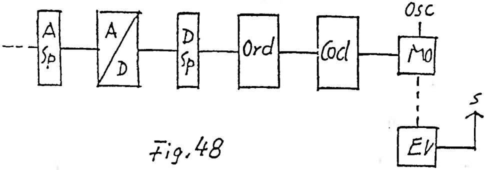


Fig. 47



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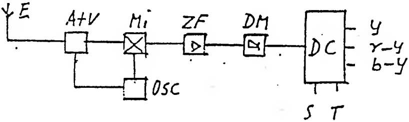


Fig. 49

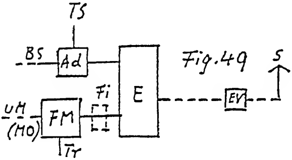


Fig. 50

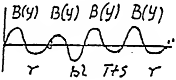


Fig. 51

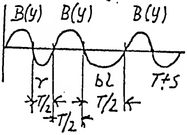
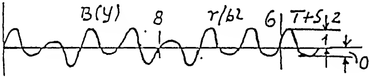


Fig. 53



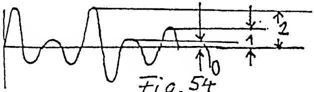
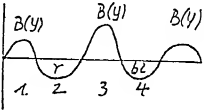


Fig. 54

Fig. 52



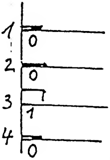


Fig. 55

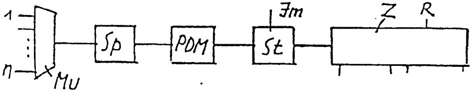
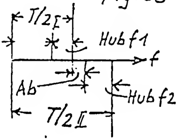


Fig. 56

Fig. 58



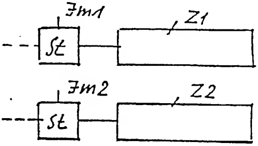


Fig. 57

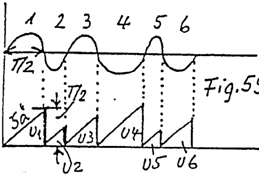


Fig. 59

Fig. 60

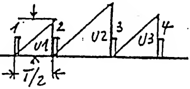


Fig. 61

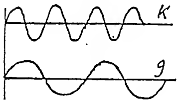


Fig. 62

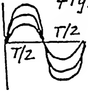
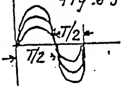


Fig. 63



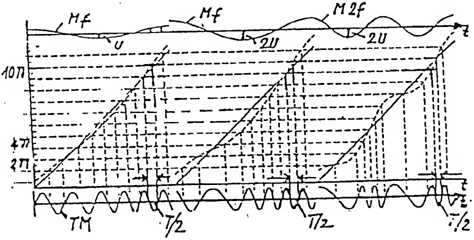


Fig. 64

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(54) **Verfahren für die digitale und/oder analoge Codierung von Information eines, zweier oder mehrerer Kanäle und/oder Frequenz- oder Bandbreitenreduzierung und/oder Erhöhung der Übertragungssicherheit.**

(57) Diesbezüglich ist bisher bekannt eine frequenz- oder zeitmultiplexe Zusammenfassung von Kanälen. Allerdings ist hierfür ein grosser Aufwand und eine grosse Bandbreite erforderlich. Bei der Erfindung werden die seriell angeordneten Codeelemente einzeln parallel geordnet und alle zusammen zu einem Codewort vereinigt. Eine Übertragungssicherheit wird in der Weise erreicht, indem die Information in PDM-Pulse umgewandelt wird und diese Impulse in die Periodendauern von Halbperioden bezw. Periodendauern umcodiert, die dann in einer ununterbrochenen Folge von positiven und negativen Halbperioden gesendet werden.

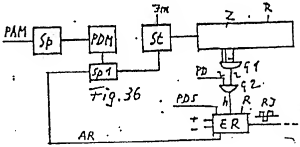
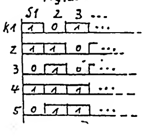


Fig.20



EP 0 329 158 A3

Europäisches Patentamt

EUROPÄISCHER RECHERCHENBERICHT

Nummer der Anmeldung
EP 89 10 2762

EINSCHLÄGIGE DOKUMENTE

Kategorie	Kennzeichnung des Dokuments mit Angabe, soweit erforderlich, der maßgeblichen Teile	Betrifft Anspruch	KLASSIFIKATION DER ANMELDUNG (Int. Cl.4)
X	US-A-4 345 323 (CHANG) * Spalte 1, Zeile 57 - Spalte 2, Zeile 9; Zusammenfassung *	1,2	H 04 L 25/48 H 04 L 27/00
X	PATENT ABSTRACTS OF JAPAN, Band 10, Nr. 337 (E-454)(2393), November 14, 1986; & JP-A-61 141 230 (SUMITOMO ELECTRIC IND. LTD.) 28-06-1986 * Zusammenfassung *	1,2	
X	US-A-4 066 841 (YOUNG) * Zusammenfassung; Spalte 1, Zeilen 31-50 *	3	
X	FR-A-2 015 695 (IBM) * Seite 2, Zeilen 14-29; Seite 10, Zeilen 11-26 *	3,4	RECHERCHIERTE SACHGEBIETE (Int. Cl.4) H 04 J H 04 J
X,P, L	EP-A-0 284 019 (DIRR) * Anspruch 1 *	3	
X,P, L	DE-A-3 802 088 (DIRR) * Anspruch 28 *	3	

Recherchenort	Abschlußdatum der Recherche	Prüfer
DEN HAAG	16-01-1990	VEAUX

KATEGORIE DER GENANNTEN DOKUMENTEN
X : von besonderer Bedeutung allein betrachtet
Y : von besonderer Bedeutung in Verbindung mit einer anderen Veröffentlichung derselben Kategorie
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Europäisches
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	GEBÜHRENPFLICHTIGE PATENTANSPRÜCHE

Die vorliegende europäische Patentanmeldung enthielt bei ihrer Einreichung mehr als zehn Patentansprüche.

☐ Alle Anspruchsgebühren wurden innerhalb der vorgeschriebenen Frist entrichtet. Der vorliegende europäische Recherchenbericht wurde für alle Patentansprüche erstellt.

☐ Nur ein Teil der Anspruchsgebühren wurde innerhalb der vorgeschriebenen Frist entrichtet. Der vorliegende europäische Recherchenbericht wurde für die ersten zehn sowie für jene Patentansprüche erstellt für die Anspruchsgebühren entrichtet wurden,

nämlich Patentansprüche:

☐ Keine der Anspruchsgebühren wurde innerhalb der vorgeschriebenen Frist entrichtet. Der vorliegende europäische Recherchenbericht wurde für die ersten zehn Patentansprüche erstellt.

X	MANGELNDE EINHEITLICHKEIT DER ERFINDUNG

Nach Auffassung der Recherchenabteilung entspricht die vorliegende europäische Patentanmeldung nicht den Anforderungen an die Einheitlichkeit der Erfindung; sie enthält mehrere Erfindungen oder Gruppen von Erfindungen, nämlich:

1. Patentansprüche 1-4,6,7,11: Verfahren zur Codierung und Übertragung von Information.
2. Patentanspruch 5: Verfahren zur Auswertung von Abständen z.B. zwischen Pulsen.
3. Patentansprüche 8,12: Verfahren zur Übertragung von Farbfernsehsignalen.
4. Patentansprüche 9,10: Verfahren für die Codierung der Farbfernsehsignale.

☐ Alle weiteren Recherchengebühren wurden innerhalb der gesetzten Frist entrichtet. Der vorliegende europäische Recherchenbericht wurde für alle Patentansprüche erstellt.

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nämlich Patentansprüche: 1-4,6,7,11

P-10568/P-11166/P-11667
EP 0329158 A (独文)の
英文翻訳

English translation of EP 0 329 158

Method for the digital and/or analog coding of information of one, two or more channels and/or frequency or bandwidth reduction and/or enhancement of the transmission reliability.

In this regard, frequency or time division multiplex combination of channels has been known heretofore. However, this necessitates great complexity and a large bandwidth. In the case of the invention, the serially arranged code elements are ordered individually in parallel and all of them together are combined to form a code word. Transmission reliability is achieved by the information being converted into PDM pulses and these pulses being recoded into the period durations of half-periods or period durations which are then transmitted in an uninterrupted sequence of positive and negative half-periods.

Method for the digital and/or analog coding of information of one, two or more channels and/or frequency or bandwidth reduction and/or enhancement of the transmission reliability.

The present invention is concerned with a method for the digital and/or analog coding of information of one, two or more channels and/or frequency or bandwidth reduction and/or enhancement of the transmission reliability.

For the transmission of information of a plurality of channels via one path, frequency and time division multiplex methods such as, e.g. the carrier frequency technique and pulse code modulation have been known heretofore. One disadvantage of these methods is that they require large bandwidths and great complexity.

The object of the present invention is to transmit the information of one, two or more channels with less bandwidth and to transmit the information of two or more channels via one channel with less bandwidth than would be necessary for the sum of the individual channels. This is done by the synchronously or quasi-synchronously arranged code elements of the different channels being ordered in parallel and all of them together being combined to form a code word and being transmitted. In addition, the intention is also to enhance the transmission reliability. This is done by the PAM pulses being converted into PDM, PPM and PFM pulses into sinusoidal half-periods or period pulses or code elements which are transmitted in an uninterrupted sequence of positive and negative half-periods. In this case, the half-period duration or period duration is a measure of the PDM-PPM and PFM pulses.

The invention can be employed, e.g. for combining telex, teletext, fax and digital telephone data

channels. The invention can advantageously be used in shared lines and line concentrators as well.

Furthermore, the invention exhibits possibilities for advantageously coding new television technologies for the improvement of C-MAC, D-MAC, D2-MAC, etc. Furthermore, it can also be used in the further development of the HDTV method. The possibilities for all these new television methods are highly restricted due to a bandwidth deficiency.

The invention is explained in more detail below with reference to drawings, in which:

Figure 1 illustrates the principle of a code division multiplex arrangement

Figure 2 illustrates the previous generation of phase jumps, e.g. in the case of 4 PSK

Figures 3 to 8 illustrate the generation of phase jumps

Figure 9 illustrates the generation of amplitude steps

Figures 10, 11 and 13 illustrate a representation of dual QAM and a vector diagram of higher-value coding

Figure 14 illustrates a vector diagram of dual QAM

Figure 16 illustrates the arrangement of the coding points in multi-value coding by means of amplitude magnitudes and phase angle

Figure 15 illustrates an overview of the generation of phase and amplitude steps

Figure 17 illustrates the generation of phase jumps

Figures 18, 19, 20, 21, 24, 28 illustrate code division multiplex examples

Figures 22, 23 illustrate an overview of a television transmitter and receiver

Figures 25, 26, 27 illustrate duplex traffic via lines and radio with just one alternating current with phase adjustment

Figure 29 illustrates the compensation of overlaps

Figures 30, 31, 32 illustrate the generation and conversion of PDM pulses into half-period pulses

Figures 33 to 38 illustrate the generation and conversion of PDM pulses into an alternating current

Figures 39 to 44 illustrate instances of coding in accordance with the invention for television

Figures 45, 46, 62, 63 illustrate a double binary and double duobinary arrangement of code elements

Figures 47, 48, 49 illustrate circuit overviews for television

Figures 50 to 55 illustrate instances of coding of colour television signals

Figures 56, 57, 58 illustrate the multiple utilization of transmission paths of PDM-coded signals

Figures 59, 60 illustrate the evaluation of phase-modulated signals

Figure 64 illustrates a graph showing the dependence of the frequency-modulated oscillation on the amplitude and frequency of the modulation oscillation.

A simple way of realizing phase jumps is described in Figures 3, 4, 5, 6 and 7. In the first instance, this will be explained in more detail with reference to Figure 3. Square-wave pulses having a frequency of 1 MHz are turned on at the transmitting end S. If, as illustrated in Figure 3c, a low-pass filter TP of 5.5 MHz is connected into the transmission path, what is almost still a square-wave pulse is obtained at the receiver E. If, as illustrated in Figure 3b, a low-pass filter TP of 3.5 MHz is connected in, the vertical edge steepness is no longer present; if, on the other hand, as

illustrated in Figure 3a, the low-pass filter is reduced to 1.5 MHz, then a sine-like alternating current having the period duration of the square-wave period is obtained at the receiver E. Thus, since the period duration does not change relative to the square-wave pulse, by altering the period durations of the square-wave pulses it is also possible to change the phase and/or frequency of the sinusoidal alternating current illustrated in Figure 3a. Since such a change always takes place at the zero crossing, a continuous change takes place and harmonics are hardly generated, that is to say the transmission is more narrowband than in the case of the phase keying systems that have been customary hitherto. At the receiving point, the change in the period duration can then also be provided as a measure of the phase jump. Such an evaluation circuit will be described later.

Figure 4 illustrates square-wave pulses with different period durations T = f, T = f1 and T = f2. After an analogous arrangement according to Figure 3a, a sinusoidal alternating current with the period durations T =1/f, T = 1/f1, T = 1/f2 would be obtained at the receiving end. Since the frequency of the alternating current decreases or increases in the event of phase jumps, the frequency change corresponds to a phase jump. This is clearly revealed by Figure 2, which illustrates conventional phase keying. It is evident from this that for each phase change a frequency change is effective, but not continuously. Therefore, it is also difficult to determine the size of the phase jump from the period duration at the receiving end.

In order to keep the frequency changes and thus also the frequency band small, each phase jump can be split into steps. Figure 5 depicts this diagrammatically. In Figure 5, T/2 is the half-period duration of a pulse and corresponds to 180 degrees. This angle is divided

into 36 steps each of 5 degrees. If a phase jump of 40 degrees is intended to be produced, then the half-period T/2 is shortened 4 times by 5 degrees, and of course so is the other half-period as well. The half-period duration relative to the reference pulse is then T1/2. After the phase jump, it is possible either to leave this frequency or else to change it over to the frequency T/2 again, by providing a phase jump of 5 degrees in the opposite direction. A phase shift of 30 degrees would then still be present relative to the reference phase. In Figure 6, the periods of the reference phase are illustrated 4 times with respect to time and the periods of the periods shortened by 2×5 degrees are illustrated 4 times. Upon comparison after the 4th period, the difference of 40 degrees relative to the reference phase is evident.

Figure 7 illustrates a circuit of an embodiment of the invention. It is assumed that the period duration is subdivided into 72 steps, to be precise with phase jump steps of 5 degrees. Each step is intended to be assigned 10 measurement pulses, and so 72×10 = 720 measurement pulses are necessary for the period duration and 360 measurement pulses are necessary for the half-period duration. Only the half-periods ever need be coded at the transmitting end. The 2nd half-period is then controlled in each case by means of the coder Cod. If phase jump steps of 5 degrees are provided, then 350 measurement pulses are necessary for the half-period, if the change is intended to be leading, and 370 measurement pulses are necessary in the case of a lagging phase change. The counting element Z in Figure 7 must therefore have at least 370 outputs. The measurement pulse frequency thus depends on the coding frequency. In the example of Figure 7, the control alternating current for the measurement pulses is generated in the oscillator

Osc. As a result, the counting element can be controlled directly via the gate G1, or, alternatively, pulses can be generated by means of a Schmitt trigger or another circuit and the counting element Z can then be switched by these pulses. The pulse duration can also be changed by altering the oscillator frequency. Assuming that the output Z2 at the counting element Z marks 370 measurement pulses, that is to say the lagging phase shift, then the coder Cod applies a potential via g2 to one input of the gate G2 which is such that, upon reaching the counting element output Z2, via which e.g. the same potential is then applied to the other input of G2, the potential at the output of G2 then changes, e.g. from h to l. In the electronic relay ER, this results in the positive potential + being applied to the output J. The coder Cod is connected to the electronic relay ER via the connection A. In the event of the next overflow of the counting element Z to Z2, ER is controlled via the connection A in such a way that negative potential - is applied to the output J. Bipolar square-wave pulses can thus be tapped off at the output of ER. Unipolar square-wave pulses could be generated in exactly the same way. This operation is repeated as long as the coder Cod applies potential to G2. If, by way of example, 5 phase steps are provided for a phase jump, then the counting element Z is switched 10 times to Z2. At the output Z2, the switch-back of the counting element is effected via the gate G4, R. Thus, by way of a differing number of outputs at the counting element Z and/or by altering the oscillator frequency, it is possible to set the pulse duration, the number of steps and the size of the steps. This variant is controlled by means of the coder Cod. The oscillator frequency can be changed over by way of fA, the number of steps and, if appropriate, the phase angle change and the step size by way of the terminals g2,

g3,..., and the amplitudes of the square-wave pulses J by way of A. Two sizes + (A) +, - (A) - are provided in the example. The square-wave pulses J are then connected to a low-pass filter in an analogous manner to Figure 3, and are passed via a transformer Ü, e.g. onto the transmission path, if appropriate with the interposition of a filter Fi.

Beginning potential must also be applied to the gate G1, via B, in order that the oscillator pulses take effect. The following instances of coding are thus possible with this arrangement: a leading phase shift, a lagging shift phase, no phase shift. These can also be effected in a step-by-step manner. The phase difference or the reference phase can be used. In addition, it is possible to provide amplitude coding, if appropriate in a step-by-step manner. A further possibility is to perform the coding upon the positive or negative pulse or half-cycle. The number of square-wave pulses is also a further code means.

It is also possible to filter out a harmonic of the square-wave pulses. If this is done, e.g. for the 3rd harmonic, then 3 periods are contained in a positive-negative pulse. The phase shifts are also contained in these 3 period durations when the pulse duration is altered.

In a wide variety of circuits, such as, e.g. in the case of quadrature amplitude modulation (QAM), alternating currents which are phase-shifted by 90 degrees with respect to one another are required. Figure 8 illustrates a circuit principle for generating such phase-shifted alternating currents of the same frequency. In an analogous manner to Figure 7, the counting element Z is controlled by an alternating current which is generated in the oscillator Osz and is passed via the gate G, at whose other input a beginning

potential B is present. In the example, the intention is to generate 4 square-wave pulses which are phase-shifted by 90 degrees with respect to one another. If the counting element Z has 100 outputs, then electronic relays ER1 to ER4 analogous to the ER relay in Figure 7 should be connected at the 25th, 50th, 75th and 100th outputs. Square-wave pulses are then generated by means of these electronic relays, as already described in Figure 7. In this case, the ER relays also contain means which always perform a potential reversal in the case of bipolar square-wave pulses and withdraw the potential during one sweep in the case of unipolar square-wave pulses. The square-wave pulses are then (designated by J in Figure 7) transmitted via the filters Fi1 to Fi4. The resultant alternating current has a phase shift of 90 degrees in each case relative to the current generated by the next output. Instead of phase-shifted alternating currents, pick-offs of e.g. PAM samples which are phase-shifted by 90 degrees can also be controlled by the outputs. A filter Fi0 is additionally arranged at the electronic relay ER1 and allows e.g. only the 3rd harmonic of the square-wave pulse to pass, with the result that the tripled frequency of the square-wave pulses is obtained here. The phase shift is then transferred to the 3rd harmonic.

With Figure 7, different amplitude steps can also be generated simultaneously. Only two are identified in the circuit. In Figure 9 there is a further possibility for generating different amplitude steps. The alternating current generated in Figure 7, for example, is fed to a limiter in which the control pulses are generated. The characteristic states are fed in via the terminal Code, which states perform a changeover to the amplitude magnitude determined by the code, to be precise in the coder Cod. The changeover to another amplitude

magnitude always takes place at the zero crossing. The magnitude of the amplitudes is determined by the resistors R1 to R4 arranged in AC circuits. Electronic relays I to Ives controlled by the coder Cod connect the various resistors into the AC circuits. Four amplitudes of different amplitudes are then obtained at the output A.

It is also known to code an information item by means of the half-cycles or periods of an alternating current; in the case of a binary code, the characteristic states are then a large and a small amplitude value. If 2 of such coding alternating currents of the same frequency are phase-shifted by 90 degrees and added, then they can be transmitted with an alternating current of the same frequency. Figures 10a,b illustrate the channels K1 and K2, which are coded by the periods as code elements with the characteristic states of large amplitude value = 1 and small amplitude value = 0. If one is phase-shifted by 90 degrees with respect to the other, then they can be added. Figure 11 illustrates their vector diagram. The channel K1 has the vector K1 (u) and the channel K2 has the vector k2 (v). The two characteristic states of the two alternating currents are designated by u1/uo and v1/vo. If both are then added, the 4 sum vectors I, IV and II, III are obtained. It can be seen that the vectors II and III no longer lie on the 45 degrees line. This makes the evaluation somewhat more difficult. Four possibilities which can all be placed on the 45 degrees line, designated by (II) and (III) in Figure 11, suffice for the evaluation of the binary signals. Figure 13 illustrates the 4 possibilities, 00, 11, 10, 01. If all 4 possibilities are on the 45 degrees vector, as illustrated in Figure 11, they can be coded by 4 amplitudes of different magnitudes, that is to say with a sinusoidal alternating current. Figure 9 illustrates

one such possibility. In order to transmit binary signals of two channels, therefore, a multi-value quaternary code is sufficient; such as e.g. 4 PSK or 4 QAM. These instances of coding are distributed between a period. In Figure 9, the positive and negative half-cycles are of the same magnitude; in that case, the transmission exhibits freedom from direct current. The positive and negative half-cycles can be utilized as an additional criterion. The 4 amplitude characteristic states can then be distributed, 2 to the positive half-cycle and 2 to the negative half-cycle. These may have the same magnitude, that is to say e.g. I + IV for the positive and negative half-cycles in Figure 11. To ensure that this coding alternating current always lies above the interference level, the coding alternating current must always have a specific magnitude, e.g. (III) as in Figure 11. The amplitude magnitude IV will than be increased somewhat.

Reduction of e.g. binary-coded alternating currents with the half-cycles or periods as code elements is already known. This presupposes phase shifts of the samples taken. The present invention demonstrates a further possibility for reducing the frequency of binary-coded information, in particular. Figure 1 depicts a channel K with a binary code 1,0,1,1.... If the frequency of the channel is intended to be reduced, into 2 channels at half the frequency, then in each case 2 serially arranged binary values of the channel K must be distributed in parallel between the channels Kv1 and Kv2; taking the example of the 4 values 1,0,1,1 of the channel K, the value 1 to Kv1, the value 0 to Kv2, the value 1 to Kv1 again and the other value 1 to Kv2. In this case, one value can always be stored, or the values can also be transmitted in a manner staggered over time. This must be taken into consideration during the evaluation. Simultaneous transmission of 2 channels has actually

already been explained in Figures 11 and 13. As is evident from Figure 13, 4 combinations are possible.

Figure 10 illustrates 4 coding alternating currents K1-K4 with the code elements of period and the characteristic states of large and small amplitude values of the same frequency. If there is a desire to transmit all 4 on the basis of QAM, they must have the following phases, K1 = 0 degrees, K2 = 90 degrees, K3 = 90 degrees and K4 = 180 degrees. K1/K2 and K3/K4 are combined to form a coding alternating current in accordance with Figure 9 and added. Figure 14 illustrates the vector diagram for this. It can be seen that 16 combinations are possible. Furthermore, it is evident from this that only 4 values lie on the 45 degrees vector. During the evaluation, the leading and lagging phase shifts must also be taken into account for the other values. The phase-shifted alternating currents are generated in an arrangement of the kind illustrated in Figure 8 and fed to 2 arrangements according to Figure 9, these alternating currents being phase-shifted by 90 degrees with respect to one another.

It is also possible to add an aggregate alternating current and single coding alternating current; a prerequisite is a phase shift of 90 degrees with respect to one another. Eight combination possibilities arise in this case.

Four channels can also be transmitted in coding division multiplex, as illustrated in Figure 1. In the first place, 16 combinations are necessary. Known instances of coding can also be provided for this, such as e.g. 16 PSK, 16 QAM and 8 PSK. Coding in this case requires one period in each case, if phase shifts in accordance with the present invention are provided. Instead of the characteristic states that are indeed in close proximity in the case of dual QAM according to

Figure 14, it is also possible to perform any desired coding. In Figure 16, the coding is performed by phase differences of 30 degrees and by 3 and 4 amplitude steps. If there is a desire for even greater reliability, the 4 amplitude steps BPh can be additionally divided. Steps may additionally be accommodated on the zero line. It is thus possible to provide each half-cycle for such coding. However, if there is a desire to perform transmission via wire-based transmission paths, it is expedient to transmit the negative half-cycle with the same coding, in order that freedom from direct current is manifested. A reduction can also be performed using the same method. In Figure 1, the channel is intended to be transmitted only at a quarter of the frequency. In each case 4 serially arranged binary elements 1 and 0 are arranged in parallel, as provided in Figures 1 a,b. The values 1,0, 1,1 of the channel K are then divided in parallel between the channel Kv1 "1", channel Kv2 "0", channel Kv3 "1" and channel Kv4 "1". In the coder, the predetermined coding point is then determined for the respective combination and transferred to the phase and amplitude of the coding alternating current. The phase is defined in Figure 7; if appropriate, it can also be used simultaneously to code the amplitude, and the required amplitudes can then be coded in Figure 9. The overview of this is illustrated in Figure 15. The coding point is defined on the basis of the four-element combination in the coder Cod. The phase coder generates the half-cycles or periods with corresponding phase and the amplitude coder generates the associated amplitudes. A phase coder may be embodied analogously to Figure 7 and an amplitude coder analogously to Figure 9.

A phase jump always signifies a change in the period duration. This change, that is to say frequency change, can be maintained if there is no further phase

change, or a changeover back to the original frequency can be effected during the next period or half-period. Since the alternating current has a different phase in the latter case, a reference phase is necessary during the evaluation. As emerges from Figure 4, with the aid of the circuit of Figure 7 it is possible to maintain any desired phase, that is to say maintain the frequency which was produced during the phase change. The phase changes are always performed at the zero crossing in the present case. In Figure 16, it is possible to provide a reference phase BPh, from which a phase shift is performed leading and lagging 2×30 degrees.

Figure 17 illustrates the generation of the phase jumps of Figure 16 according to the principle of Figure 7. The angle of 360 degrees is identified by 3600 pulses. If there is only an amplitude change with the reference phase, then the counting element is always switched through from 0 to 360 degrees. In this case, the control is effected by means of the coder Cod, which has already been described in Figure 7. In this case, the amplitude change is effected in the manner illustrated in Figure 7 or in Figure 9. If the phase jump Ph1 in Figure 16 is intended to be effected, then it is necessary, if freedom from direct current is required, for each half-period to be switched as far as the output 195. A reference phase is not necessary during the evaluation because, as long as no further phase change takes place, the unambiguous phase is, after all, defined by the period duration. If the coding lies on the vector Ph3, then the period duration is 330 degrees, that is to say a changeover is always effected at the output 165. In this case, the phase shift is always referred to the period duration. If, e.g. in the last case, the phase shift were referred to the half-period, then a switch-back would in each case have to be effected at the output

150. Other methods of generating phase jumps can be used in exactly the same way.

The phase jumps are evaluated in a known manner by measuring the period durations by means of an excessively increased control rate of counting elements, e.g. disclosed in European patent application 86104693.6.

A reference phase is necessary in the evaluation of Figure 14. The amplitude points 1 to 4 are arranged directly on the reference phase angle while the other 12 coding points are arranged such that they are leading and lagging with respect to the reference phase. It is assumed that the signals are those of a television system. In the blanking interval, the reference phase is then determined and, at the same time, control signals are transmitted. In this case, only the amplitude values on the reference phase are used. From the transmission path ÜW, the signals are fed to the input unit EST (Figure 12). They then pass to a limiter B, on the one hand, and to a code evaluation arrangement CA, on the other hand. In the limiter, the positive and negative half-cycles are converted into Jp and Jn pulses. In the comparison device VE, the phase of the pulses arriving from the transmission path is then compared with a reference phase pulse JBn. Figure 12a illustrates the leading, lagging and reference phase pulses Jv, Jn, JB which are compared with the reference phase pulse JBn determined from a coding. The 3 possible phase values of leading, lagging or reference phase are each passed to the code evaluation arrangement. In the latter, the amplitude values are determined and, in connection with the leading, lagging or reference phase, the coding points are then determined and forwarded via S for further utilization. The coding of the reference phase in the blanking interval may be configured e.g. in such a way that the point 2 is transmitted for 4 times and the

point 4 is transmitted 4 times on the reference phase. The evaluation thereof is carried out in the reference phase evaluation arrangement BA. The latter then passes a reference phase pulse JBn to the comparison device.

Figure 18 illustrates a further exemplary embodiment of the invention. The 5 channels K1 to K5 are intended to be transmitted by code division multiplex only via one channel or path. The e.g. binary-coded information of these 5 channels is firstly stored in the memory Sp. By way of example, Figure 20 illustrates the steps of the binary characters, to be precise in a manner such that they are already synchronized. Therefore, in each case 5 steps or pulses S1, 2, 3,... arranged in parallel are to be coded. The steps of S1 are 1-1-0-1-0. Five bits are necessary for the coding of these 32 combinations. In the example, these are coded with the amplitudes of the half-cycles of an alternating current with the characteristic states of large and small amplitude values and with a leading and a lagging phase jump of 36 degrees, as is shown in Figure 19. The binary values are fed to the coder Cod from the memory Sp of Figure 18 and are converted into a corresponding code in the said coder. In the decoder at the receiving end, the corresponding steps are assigned to the 5 channels again in accordance with the code.

Figure 21 illustrates a further application of the invention for the coding and transmission of the signals in colour television. The luminance signal is tapped off at 6 MHz. This principle has actually already been disclosed in the published patent application P 32 23 312. The colours red and blue are each intended to be tapped off at 1.2 MHz, that is to say one red and one blue tapping coincide with 5 luminance tappings. The luminance tappings are designated by I, II, III, IV, V. These samples taken are coded with 8 bits, binary-coded

in the example. With the tapping III, the tappings for red and blue must then also be carried out. The samples taken of red and blue are binary-coded with 6 bits in the example. During the transmission of the 5 luminance samples taken, the code for the red and blue colour samples taken is also simultaneously transmitted. With the tapping of red and blue, the transmission of the colour and the sampling I of the luminance signal could be begun. It is also possible to store all 5 luminance samples taken and colour signal samples and begin the transmission of all the television signals only after the 5th sampling. Figure 21a depicts the binary codes of all the signals to be transmitted. The 8 bits 1-8 of the luminance samples taken are each arranged in parallel. Then, serially, digital audio and other signals T + So, the 6 bits of the red signal and once again the audio and other signals are arranged under 9, 10, and the audio and other signals again and the 6 bits of the blue signal are arranged under 11, 12. It is expedient if the luminance samples I to V are still stored at the transmitter and the colour codes for red and blue are transmitted with the preceding luminance samples, so that it is then unnecessary for the 5 luminance samples to be stored at the receiver. Only the red and blue samples then need be stored. The audio and other signals must likewise be stored and then be fed to the loudspeaker contemporaneously with the picture. These signals can, of course, also be placed in the blanking interval. In the example, therefore, 12 bits are required for the transmission of a luminance sample for the audio and other signal samples and for the colour samples taken. Figure 21b illustrates an example of the coding of these 12 bits. Five half-periods of an alternating current are provided for this purpose. In this case, the binary code comprises code elements of the half-cycles with the

characteristic states of large and small amplitude values. In addition, a leading and lagging phase shift of 36 degrees is also provided, with the result that 12 bits are thus obtained.

Figure 22 illustrates an overview of such a television transmitter. The control element StO controls the television camera FK and also supplies the remaining control signals such blanking and synchronizing signals A + S. The red, green and blue signals are fed in the first place to the Y matrix YM and red and blue are simultaneously fed to the chrominance conditioning arrangement FA. At the same time, a concentrator K is provided, which taps off the luminance signal Y, the colour signals r + bl and the audio and other signals. At the tap 3, a criterion is passed via the connection 3a to the chrominance conditioning arrangement. In the latter, tapping off from the red and blue signals is performed and both values are stored in the capacitors C1 and C2. A Y value present at the 3rd tap is additionally fed to the FA by the Y matrix, with the result that the colour difference signals r-y and b-y are obtained at the taps 6a and 6b. - It is also possible to tap off just the colour separation signals. - By means of the module TSo, the audio and other signals are fed to the concentrator in an analogous manner via 6c and 6d. From the concentrator, all values are fed to a memory Sp. From the memory, the signals are fed in a correctly timed manner, e.g. as described in Figure 21a, to an analog/digital converter. In the latter, coding is effected in accordance with Figure 21b. During the blanking interval, a changeover is made to the concentrator K1 via U. As a blanking criterion, it is possible e.g. sometimes to transmit the code word with all zeros. --- In addition, other signals So can also be transmitted in the blanking interval. The beginning of a line can also be marked by

a zero code. Synchronization is predetermined during the line by the sequence and the number of half-cycles. In the case of the present code, a nominal frequency of 15 MHz is necessary. If there is a desire to use only one amplitude code, 2 alternating currents each at 18 MHz are necessary, which could then be phase-shifted by 90 degrees and added before being transmitted. It is merely a question of viability and reliability as to which method is used here. The leading or lagging phase jump is defined by the period duration in the example. Therefore, no reference phase is necessary in that case. It is possible, of course, to use multi-step amplitude codes and/or phase codes in order to reduce the frequency. The PAM signal, for example, can be applied to the audio input T, which signal is then tapped off occasionally within the 8 kHz time. In this case, there are numerous opportunities for utilizing the tap 6c/6d. Figure 23 illustrates a partial overview of a television receiver. The signals are fed to the demodulator DM via the RF oscillator and mixing stage and the amplifier V. In the said demodulator, e.g. the signals as illustrated in Figure 21b are obtained again and fed to the decoder DC. The colour signals are subsequently forwarded to the matrix Ma. The Y signal is also connected to the said matrix. By way of example, the colour difference signals R-Y, G-Y and B-Y are then obtained at the output of the matrix and, like UY, are passed to the television tube. The decoder DC then additionally supplies the blanking and synchronizing signals AS and the audio and other signals.

Figure 24 illustrates an example in which the code for the code division multiplex is obtained from a plurality of alternating currents. It represents a binary code in which the half-cycles of the alternating currents serve as code elements and in which a large and a small

amplitude value form the characteristic states. The characteristic identifiers to be transmitted comprise square-wave pulses at the frequency 1000 Hz, as is illustrated in Figure 24a. Twenty channels are intended to be transmitted in code division multiplex. The half-cycles of the alternating currents at 1000, 1500, 2000, 2500 and 3000 Hz are provided for this purpose. A plurality of channels at a lower bit frequency can, of course, be fed to each channel in time division multiplex. The same bit number could be achieved in exactly the same way with 2 alternating currents at 2000 Hz and once again 2 alternating currents at 3000 Hz, in which case these would each have to be phase-shifted by 90 degrees with respect to one another, so that they could be added in the event of transmission. The best way of producing synchronization between the individual channels is already known (Unterrichtsblätter der DBP Issue 4/6, 1979), and it will not, therefore, be discussed any further. Digitized voice or a plurality of voice channels can also be transmitted simultaneously in the same way.

In the case of amplitude coding, duplex operation can be carried out using the same alternating current. To that end, it is necessary for the remote coding alternating current to be phase-shifted by 90 degrees. Figure 25 illustrates this principle. In this case, the code may be digital, a binary code in accordance with the patent DE 30 10 938, or, alternatively, analog in accordance with Canadian patent 1 214 227. With half-cycles as code elements the frequency is 32 kHz in the case of digital coding and 4 kHz in the case of analog coding. In Figure 25, S1 is the microphone and E2 the receiver of one subscriber and S2 and E1 those of the other subscriber. In S1 there is also a coder in which the coding alternating current is

obtained from the speech. From S1, the coding alternating current passes via a hybrid G, the subscriber or connecting line RL to the hybrid G of the remote subscriber and to the receiver E1. The latter additionally contains a decoder which recovers the speech from the coding alternating current. The coding alternating current of S1 shall be the synchronizing alternating current. From E1, the said current is branched off via a 90 degrees phase shifter to S2, in which it is amplified, if appropriate. If S2 now speaks, a coding alternating current which is phase-shifted by 90 degrees is transmitted via G, RL, G to E2, where it is decoded and communicated to the receiver as speech. If, by way of example, simultaneous speaking occurs momentarily, an addition alternating current is produced on the transmission path RL. Cancellation is not caused. This principle can be provided in exactly the same way for duplex traffic in the case of data transmission. Further examples in this regard are disclosed in the published patent application DE 3802088.

This method can, of course, also be used for radio, e.g. for directional radio. Figure 26 depicts an overview in this regard. In this case, the transmission alternating current is concomitantly provided as the coding alternating current at the same time. Low-level modulation is advantageously used. The transmission alternating current is generated in the oscillator Osz1. The basic signal is converted into an alternating current digital code in the analog/digital converter A1/D1. It is even more simple if an arrangement according to Figure 7 is provided as oscillator and coder. From the coder, the electronic relay is then controlled in such a way that large and small square-wave pulses are present at the output J, and are then shaped to form a sinusoidal alternating current in the low-pass filter TP. The coding

alternating current then passes via amplifiers (not illustrated) to the output stage E and to the transmission antenna. A branch circuit may additionally be provided in the output stage, in which branch circuit the harmonics are phase-shifted by 180 degrees, and are then fed to the main circuit again for the purpose of compensation. At the receiving end, the useful signals are fed via a fixed tuning circuit to an amplifier V and then forwarded to the digital/analog converter D2/A2. The analog signal is then passed on, e.g. via a switching system. Via the amplifier V, the transmission alternating current is also branched off to a 90 degrees phase shifter Ph and then forwarded to the oscillator Osz2. The oscillator is synchronized with this. Via the converter A3/D3, amplifiers (not illustrated) and the output amplifier E, the transmitter of the opposite direction is then operated. The receiver E1 is connected in exactly the same way as the receiver E2, only the phase shifter is not necessary.

A phase shifter according to the principle of Figure 7 is illustrated in Figure 27. In the latter, compensation for small frequency fluctuations is provided at the same time. For this purpose, a counting element Z is provided which has 1000 outputs. During a half-cycle of the transmission alternating current, the counting element passes through these 1000 outputs. The control pulses Js are generated in an oscillator (not illustrated). In the case of a phase shift of 90 degrees, a phase shift of 45 degrees coincides with a half-cycle; that corresponds to 250 outputs. The transmission alternating current half-cycles coming from the amplifier V are fed to a limiter, with the result that square-wave pulses Jp and Jn are produced at the output thereof. These pulses are connected to the control element St. The control pulses Js and the beginning characteristic

identifier Be are additionally applied to the said control element. The control element is connected in such a way that only whole Jp and/or Jn pulses are ever activated at the counting element. If the counting element has reached the output 1000 during a pulse Jp, then the gate G11 assumes the operating position. A Jn pulse is connected to the gate G12 and, after the end of the Jp pulse, as a result of the delay of the monostable element mG4, potential is also momentarily connected to the said gate G12. G12 is activated and applies potential to one input of G13; I potential has already been applied to the other input of G13 from G11. A potential changeover then takes place at the output of G13 and inverts G16 at the output. The consequence of this is that G17 generates a switch-back potential for the counting element. Potential is also applied to the gates G8, G9 and G10 such that they, in interaction with the allocated outputs 1000, 999, 1001, control one of the monostable elements mG1, mG2 or mG3. Since the Jp pulse has controlled the counting element up to 1000, the gate G9 and mG2 has now been activated. If the counting element is then controlled to the output 250 by the next Jn pulse, then the gate G6 is activated, which controls the electronic relay ER which, in accordance with Figure 7, generates a square-wave pulse which is shaped to form a half-cycle in the low-pass filter. For the Jn pulse, the gates G15, G14 and the monostable element mG5 are arranged for the output marking. The monostable element mG2 is latched, e.g. up to the output 260. G6 then assumes the starting position again. The electronic relay remains in this position until the next marking of the output 250. If only the output 999 is reached due to a frequency fluctuation, then, instead of G9, the gate G8 is marked and mG1 and G5 are activated when the output 249 is reached. If the output 1001 is reached, then G10

and mG3 are activated, and the gate G7 is activated in the event of the output 251 being reached. Such frequency fluctuations are thus also passed on to the alternating current which is phase-shifted by 90 degrees. Figure 27a illustrates the control element in detail. The pulses Jn and also the beginning characteristic identifier are connected to the gate G3. If both are present, G3 is activated and causes the bistable element bG to attain the operating position, which then applies operating potential to the gate G1. It is only then that the Jp pulse can take effect. The control pulses Js then pass via the gate G2, which is merely a potential reversal gate, to the counting element. The further operations at the counting element have already been described.

In Figure 27, the negative half-cycle can be generated either by the Jn pulse, or the sweep of the positive half-cycle is repeated, the respectively marked outputs being stored.

The code used in the invention may preferably be an amplitude and/or phase code, of the kind illustrated by way of example in Figure 16. With purely an amplitude code, it is also possible to provide 2 code alternating currents of the same frequency, in which case one is then phase-shifted by 90 degrees in the event of transmission and subsequently added to the other.

The principle behind the invention can also be used for the transmission of digitized voice. Figure 28 illustrates 5 coding alternating currents with a binary code, the characteristic states being a large and a small amplitude value of the respective half-cycle. In this case, the frequencies are 8, 12, 16, 20 and 24 kHz. Twenty bits are obtained in this case; if 2 alternating currents of the same frequency, but phase-shifted by 90 degrees, are additionally provided, then 40 bits are obtained, that is to say, in the case of 8-bit code

words, as illustrated in Figure 28a, 5 digitized voice channels can thereby be transmitted.

In Figures 21 and 22, 2 audio tappings suffice per line given a tapping frequency of approximately 30 kHz (PAM) per line, which tappings can be effected e.g. at the beginning of the respective picture line and in the centre of the picture line; the spacing is then 32 µs. Each tapping is then converted into an 8-bit code in the analog/digital converter A/D and is then transmitted with the following 5 luminance code words, as is illustrated in Figure 21a. By way of example, with I/9, 1o, 11, 12 and V/9, 10, 11, 12 in Figure 21a. The tappings during the frame change time must be determined e.g. by time measurement. The coding is then also effected in the frame change time.

For the code division multiplex it is possible, of course, to use any desired code, such as the AMI or HDH-3 code. In the examples, an amplitude code is often used in which the code elements comprise the half-cycles or periods of a sinusoidal alternating current with the characteristic states of small and large amplitude values. In this case, one code element corresponds to one bit. If, by way of example, 12 bits are required for the CVBS and audio signals, then 12 half-cycles are necessary. The coding can be realized asynchronously with the tappings, since the length of the code words does not change. If, on the other hand, a phase code or additionally a phase code is provided, then the period duration also changes in the event of each phase change, with the result that, in the case of a periodic tapping and in the case of equidirectional phase changes, the signal tappings are no longer synchronous with the code. For compensation purposes, there are two possibilities in this case - in addition to buffer storage - in the first place re-establishing the nominal frequency in the event

of each phase change until the next phase change, e.g. in Figure 4 the nominal frequency f2 and, if a phase change T = f1 takes place and if the following codings have the same phase changes, then the following codings are coded with the nominal frequency f2. Only if the phase f1 changes again does a phase change then take place with regard to the reference phase, that is to say the reference phase must be stored at the receiver. The said reference phase can be transmitted by the transmitter, e.g. in the blanking interval. Another possibility for avoiding overlaps of 2 tappings consists in the following procedure: at the transmitter, with each code word, a measurement is made between the end of the code word and the preceding and the succeeding tapping. If there is the risk of an overlap in the leading or lagging direction, then code words having the smallest or largest period durations are interposed. Such code words are illustrated in Figures 29a and 29b. This can be circumvented by line storage.

In Figure 19, a code element has 6 different steps and the code word has 2 positions; consequently, 6 to the power of 2 combinations are possible, that is to say 36 combinations. Five bits are obtained with 32 combinations. In Figure 21b, a code element can likewise assume 6 steps, with the result that, given 5 positions, 6 to the power of 5 = 5184 combinations are possible, that is to say at least 12 bits. 4096 combinations are obtained with 12 bits.

In Figure 22, the PAM for the audio is generated in the TSO element and applied to 6c in each case, e.g. in a half-line by half-line manner. The terminals 6c and 6d are not necessary if the audio and the other signals are placed in the blanking interval, so that the concentrator K1 then performs these tasks.

The way in which e.g. the code division

multiplex can also be applied to television shall be shown with the aid of Figures 21, 22 and 23. The transmission frequency can, of course, be significantly reduced if more amplitudes and/or phase steps are provided. In addition, it is also possible to effect a combination with different carriers, as envisaged e.g. in the patent application P 32 29 139.6 Figure 9, or with different current paths. Thus, e.g. in Figure 28, a 64-kbit voice channel can be transmitted at 8 kHz, to be precise with a binary code. Two positions are each marked by the 2 half-cycles of an 8 kHz alternating current, and 2 further positions by the 2 half-cycles of an alternating current which is phase-shifted by 90 degrees. These 2 alternating currents are summed and transmitted as one alternating current via one current path. The same is carried out via a 2nd current path, so that the code word has 8 positions and 2 steps, with the result that 256 combinations are obtained. At the receiving end, decoding is performed after the evaluation of the half-cycles and, of course, buffer-storage. The coding can also be effected in a duobinary fashion.

A further method of transmitting, in a frequency-modulated manner, in particular analog signals such as voice, sounds, the luminance signal in television, the colour signals in television, telecontrol values, to be precise with less bandwidth, consists in converting the magnitude of the PAM pulses into PDM pulse lengths with the aid of pulse duration modulation PDM. These PDM pulses can then be converted into alternating current pulses, e.g. according to the method of Figure 7. The pulses are then formed by the half-cycles or periods of an alternating current, the period durations or half-period durations of the half-cycles or periods being equal to the length of the PDM pulses.

The spectrum of the frequency-modulated

oscillation used hitherto contains a large number of side oscillations above and below the carrier, which means that a very wide band is necessary in the case of transmission. In this case, the required bandwidth is greater than twice the frequency swing. In the case of the circuit according to the invention, predominantly digital switching means can be used, thereby enabling inexpensive production.

The method will now be explained in more detail below with reference to drawings. Firstly, known circuits will once again be explained, these being necessary inter alia in the context of generation (European patent application 0 284 019). Two exemplary embodiments of the invention are described below. Firstly, the principles behind the two embodiments are summarized. The information is in the first place subjected to pulse amplitude modulation and subsequently converted into pulse durations with the aid of the equidistance method, or else the information is directly coded into pulse durations with the aid of the sawtooth method. These pulse durations are then converted, in conjunction with the intervals between the pulse durations, into square-wave pulses and subsequently into sinusoidal coding alternating currents with the aid of filters. The pulse durations and intervals are converted with the aid of counting elements in conjunction with electronic switches. The pulse duration then corresponds to the duration of a half-period or period of the coding alternating current. If the pulse duration is short, the frequency of the half-cycle or period in the coding alternating current is high; if the pulse duration is long, then the frequency of the half-cycle or period in the coding alternating current is small. At the receiving end, the half-period or period durations are evaluated, for example by measurement. In this case, therefore,

frequency and phase modulation is simultaneously present.

In the case of the 2nd embodiment, the pulse duration pulse, PD1, PD2 in Figure 32, and the interval between the pulse durations (Figure 32, P) - the pulse duration and the interval each correspond e.g. to the interval between 2 tappings, designated by tp in Figure 30a - are fed to an electronic relay in which bipolar square-wave pulses are then generated. The frequency-modulated coding alternating current is then generated with the aid of filters.

Figure 7 illustrates how the time of a pulse is determined with the aid of a counting element Z in conjunction with the frequency of the stepping or measurement pulses generated in the oscillator Osc. The respective output of the counting element then marks the time. This is then provided in conjunction with gates for the control of an electronic relay ER. The latter then generates bipolar square-wave pulses.

The detailed functioning is as follows. The stepping or measurement pulses for the counting element Z are generated in the oscillator Osc. The said pulses pass via the gate G1 to the counting element Z as long as the beginning characteristic identifier is present at B. In the example, only the outputs Z1 and Z2 of the counting element are required. These outputs are connected to the gates G2 and G3. If the half-period of the square-wave pulse J is intended to have the magnitude of the sum of the measurement pulses up to Z1, h potential is applied to g3 from the coder Cod, with the result that a potential changeover takes place at the output of G3 when the output Z1 is reached, which potential changeover causes the electronic relay ER to end the square-wave pulse. If this was a positive pulse, then the next pulse will be negative. The counting element is then switched back again in this position. The

gate G4 is provided for this purpose at the output z2. From the coder, the oscillator frequency can also be increased or decreased via fA, with the result that, by way of example, different times could be marked by the respective outputs. A connection A also passes from the coder Cod to ER, and can be used to control different pulse magnitudes J.

The square-wave pulses are passed onto the line as a sinusoidal coding alternating current via a low-pass filter Tp, the transformer Ü and the filter Fi. The half-period or period of the coding alternating current is the same as that of the square-wave pulse. The principle behind the conversion of the square-wave pulses into a sinusoidal alternating current is illustrated in Figure 3. If, by way of example, square-wave pulses at the frequency 1 MHz are band-limited by a low-pass filter of 5.5 MHz, then rather steep edges are still obtained, as is illustrated in Figure 3c. A low-pass filter of 3.5 MHz was inserted in Figure 3b; it can be seen that the edge steepness has already diminished to a noticeable extent in this case. In Figure 3a, a low-pass filter of 1.5 MHz is connected in, and a sine-like alternating current is obtained at the receiver in this case. The period durations are identical to those of the square-wave pulses, that is to say that the period durations can be taken as a measure of the frequencies and/or phases. This principle was used in Figure 7 in the conversion of the square-wave pulses J into a coding alternating current with the aid of the low-pass filter TP.

Figure 4 depicts square-wave pulses having different period durations, to be precise expressed by the frequencies f, f1 and f2. These square-wave pulses have mutually different phase shifts and/or different frequencies. It can be seen from this that phase jumps and/or frequency jumps can be caused by changing the

period durations, so that frequency modulation is also obtained by this means. In Figure 5, such a phase and/or frequency jump is effected in a step-by-step manner. What is achieved as a result of this is that the bandwidth becomes small. As revealed by Figure 6, given phase jumps of 5 degrees per 180 degrees, a total phase shift of 40 degrees is obtained in the case of 4 phase jump steps.

Figure 30a illustrates PAM-coded pulses of a signal Inf. These pulses are converted into pulse duration pulses with the aid of an equidistant method, as is shown in Figure 30b. The distance between the PAM pulses (Figure 30a, tp) corresponds in each case to a pulse duration PD and an interval P, as illustrated in Figure 30b. Pulse duration modulation can also be carried out with the aid of the sawtooth method. This method is illustrated in Figures 31 and 32. The pulse durations are square-wave pulses PD1, PD2. Symmetrical PDM and bipolar PDM are also known (also see the book "Modulationsverfahren" [Modulation methods] by Stadler 1983).

Figure 35 illustrates an exemplary embodiment in accordance with the invention. In the pulse duration modulator PDM, the pulses are generated, e.g. according to Figure 30b or 32, and are passed via G5 to the gate G1. The measurement pulses Jm, e.g. at a frequency of 100 kHz, are present at the other input of the gate G1. As long as a PD pulse is present at G1, the measurement pulses Jm are activated at the output. The measurement pulses pass via the potential reversal gate G2 to the counting element Z, which is controlled by these pulses. The number of outputs at the counting element corresponds e.g. to the distance between two PAM pulses, tp in Figure 30a. Suppose that the tapping frequency is 10 kHz; the counting element would then have 100 000 outputs. The frequency swing is determined by the largest and smallest

amplitude values of the information item Inf, designated by gw and kw in Figure 30a. The outputs of the counting element Z lead to gates G3 and the outputs of the gates lead to gates G4. The respective PD pulse is present in each case at the other input of the gate G4, which pulse inhibits the gate G4. Only when the PD pulse is no longer there can the output potential also be activated at G4 via G3. ER then receives via G4 a potential changeover characteristic identifier for the next square-wave pulse. The beginning of the square-wave pulse is marked by the respective PD pulse. The next square-wave pulse is determined by the interval P (Figure 30b, P). From ER, a potential is applied to gate 5 via P, in order that the measurement pulses Jm become transmissive again at the gate G1. The counting element Z is then switched up to the output for gate G6. When the next PD pulse arrives again, G6 is activated and the counting element is switched back to the starting position via R. At the output of ER there are then square-wave pulses RJ having the magnitude of the half-periods like that of the PD pulses and of the intervals P. In the filter Fi, the square-wave pulses become sinusoidal half-cycles fmo, and so the information is frequency-modulated. The half-periods of the useful signal modulation frequencies then vary between the half-period durations identified by kw and gw at the counting element. In Figure 33, by way of example, kw = 15 kHz, the centre frequency is 10 kHz, and, in Figure 34, gw = 75 kHz. In the example, the pulse durations may change by half; this is a dimensioning matter of the pulse duration modulation circuits. The half-cycles of the intervals have a minimum frequency of 7.5 kHz in Figure 33 and a maximum frequency of 15 kHz in Figure 34. The amplitudes of the half-cycles always remain the same. The evaluation at the receiving end is effected by measuring the half-period durations.

Synchronization is not necessary since the zero crossings of a period simultaneously code the tappings in the case of coding with the aid of PAM; therefore, only the positive half-cycles need be converted into PAM pulses. The PAM pulses are then lagging by a period at the receiving end.

The redundancy of the intervals in Figure 35 can be avoided if, by way of example, the PAM pulses are stored and the next PAM pulse is called up after each PD coding. However, synchronization is then necessary at the receiver. If PAM were used at the transmitting end, the tapping frequency would have to be synchronized from time to time. Figure 36 illustrates the basic circuit of such a circuit at the transmitting end. The PAM pulses are stored in the memory Sp. The call-up of the next pulse arrives from ER via AR. In preparation, the next pulse had already been stored as PDM pulse in the memory Sp1. As a result, the counting element Z is then controlled by means of the control element St and set to a corresponding output. The counting element has also been returned to the starting position by ER via R. The control pulses Jm are also present at the control element. With the call-up of the PDM pulse, a PAM pulse is also passed from the memory Sp to the pulse duration modulator and is stored in the latter as a PDM pulse until the Sp1 memory is free again. Two Sp1 memories will expediently be provided and will then be connected to the control unit alternately after each call-up by ER. At the end of the PDM pulse, an end-of-criterion is passed to ER via the counting element Z, G1, G2. The square-wave pulse PD generated by ER is inverted to the next one, the counting element is switched back via R and, via AR, the call-up of the next [lacuna]

Figure 39 illustrates 4 channels with half-cycle coding with the characteristic states of large and

small amplitude values. The frequency is the same for all 4 channels. These 4 channels are provided for coding the colour television signals. Eight bits are provided for the Y signal (luminance signal), to be precise in each case 4 bits for the channels a and b; in each case 2 bits in the channels a and b are provided for audio and other signals T + S. The channel c is present for the coding of the red signal and the channel d is present for the coding of the blue signal, with 6 bits in each case. In each case 2 channels are then combined in accordance with Figure 11 vector I, (k1, k2) with the instances of coding I, (II), IV, (III), thereby resulting in an aggregate alternating current in accordance with Figure 9. The phase angle of the two aggregate alternating currents is then fixed at 0 degrees and 90 degrees. These 2 aggregate alternating currents can then be transmitted on the basis of quadrature amplitude modulation, with the result that a narrow band is required for transmitting all the colour television and other signals. Transmitted as dual QAM, that is to say channel a + b quadrature-amplitude-modulated and channels c + d quadrature-amplitude-modulated, where the channels have phase angles of 0°, 90°, 90° and 180° with respect to one another and their aggregate alternating currents have phase angles of 45° and 135°, and where the two aggregate alternating currents are again subjected to quadrature amplitude modulation, the evaluation is more difficult, as is also evident from Figure 11 (the vectors I, II and II are produced in the case of single QAM).

The 4 channels or their binary values can also be transmitted in code division multiplex. The binary values of the 4 channels are illustrated once again in Figure 40. In accordance with Figure 41, in each case 2 rows of Figure 40 are intended to be combined into 8 bits. In Figure 39, suppose that the frequency of the

alternating currents is 6 MHz; 18 MHz are then required for the coding. If, in Figure 41, use is made of duobinary coding in accordance with Figure 62 with the half-cycles as code elements, then although there would be a slight gain in bandwidth relative to Figure 39, the frequency would be 3 times as high. If the rows 1, 2, 3 and 4, 5, 6 are combined, that is to say 12 bits in each case, in this duobinary code, then a code word having 3 steps and 8 positions is necessary for one row 1, 2, 3. Eight positions mean 4 periods. A frequency of 2×24 MHz would thus be necessary, that is to say also too high for this purpose. Figure 45 illustrates a code element having 4 steps. With 4 steps, this results in 256 possibilities. Coding according to Figure 41 would result in a frequency reduction to 36 MHz. Figure 63 illustrates a code element having 6 steps. In order to serially code 3 rows of Figure 40, that is to say 12 bits, 5 positions would be necessary here. 30 MHz would thus still be necessary. In addition to the 3 amplitude steps, 2 phase steps or period durations are also provided. Figure 46 illustrates 3 amplitudes and 3 phase steps. If 2 rows each of 12 bits are formed from the arrangement of Figure 40, 3 positions are necessary for each row, that is to say 6 positions for both rows, in other words a frequency of 18 MHz is necessary.

The colour television signals are arranged differently in Figure 43. Eight bits for a Y tapping (luminance, pixel B) are serial each with 4 bits, and the colours red or blue are serial each with 3 bits in the rows III + IV. The respective 4th bit in rows 3 and 4 is provided for audio and other purposes. The colour red or blue respectively appears with every 2nd Y signal, that is to say these continually alternate. If the vertical rows 1/2 and 3/4, as illustrated in Figure 44, are combined, then more favourable conditions result in the

event of coding. With 4 steps, 3 positions are necessary; a frequency of 18 MHz is then necessary. If the rows 1/2 and 3/4 are arranged in parallel, that is to say 16 bits, then 4 positions, that is to say a frequency of 12 mHz, are necessary in the event of coding according to Figure 46. The dual QAM of Figure 39 can be transmitted in a frequency-modulated manner in order to provide even more reliability during transmission. The aggregate alternating current has only small frequency changes, with the result that, as revealed by Figure 64, the frequency-modulated oscillation can indeed be transmitted in a narrowband fashion. This figure reveals that the half-period duration T/2 becomes very short in the event of a frequency increase, in other words that the frequency greatly increases. With a modulation frequency Mf and an amplitude u, the half-period duration is T/2; with a doubled amplitude 2u, the half-period duration is shorter, while with the frequency doubled in addition, frequency M2f, the half-period duration is substantially reduced.

Figure 47 illustrates an overview of a television transmitter in which the codes explained in Figures 40, 41, 43 and 44 are used. From the multiplexer (not illustrated) the analog signals that have been tapped off arrive and pass into the analog memory ASp, from where the samples taken are forwarded to one or more analog/digital converters. The digitized signals are then stored in the digital memory DSp and subsequently fed to the ordering unit. In the latter, they are ordered in accordance with Figure 40, 41, 43 or 44. Having been ordered in this way, they are fed to the coder. They are coded in accordance with the predetermined code, e.g. according to Figure 45 or 46 or 62 or 63, and fed to the modulator MO. The transmission alternating current is fed to the modulator from the oscillator and the modulated

transmission alternating current is passed via amplifier stages (not illustrated) and the output amplifier to the antenna. An overview of the receiver for evaluating the coded signals is illustrated in Figure 48. A transmission alternating current arrives via the reception antenna E and passes into the stages tuning circuit/amplifier, mixing stage/oscillator Mi/Osc, via the intermediate frequency amplifier ZF to the demodulation stage - the input is connected like a superheterodyne receiver in the case of broadcasting reception -; the code alternating current is present at the output of the demodulator. The said current is connected into the decoder. The signals tapped off in the transmission multiplexer are obtained again here, such as the Y, r-y, b-y, audio and other signals S, and fed to the various circuits.

Figures 50 and 51 illustrate instances of analog coding of the colour television signals. An alternating current of the same frequency as the code alternating current is provided in Figure 50. The amplitudes of the half-cycles are the code elements. The tapping sequence is y, r, y, bl, y, T + S, etc. These analog-coded signals are transmitted on the basis of frequency modulation, with the result that a narrowband - only one frequency Figure 64 - and also transmission reliability are obtained.

An analog code is likewise provided in Figure 51. It is phase coding. The analog code is manifested by half-period durations of different lengths. In this case, the amplitudes of the half-cycles always have the same magnitude; it is a kind of frequency and phase modulation. The individual signals are arranged serially again, in the example y, r, y, bl, y, T + S. The transmission is effected at 6 MHz given a tapping frequency of the Y_ signal at 6 MHz. If multiplex tapping of all the signals is effected, that is to say including

the r, bl and T + S signals, then a tapping frequency of 12 MHz is necessary.

Coding in accordance with Figure 51 is provided in Figure 52, except that the audio and other signals T + S are coded by a superposed amplitude code. It is a binary code with a large and a small amplitude. The values of the Y and r + bl signals are defined by the half-period durations. In synchronism with the PDM pulse, the respective amplitude value is then passed e.g. to the ER relay of Figure 36, in which a square-wave pulse with a small or large voltage is then generated. The amplitude code elements may, for example, be assigned to a plurality of channels, such as audio stereo, etc. In Figure 55, the 4 half-cycle code elements are assigned to 4 different channels.

An evaluation of the PDM, PPM or PFM pulses with the half-period durations coded is evident from Figure 59. This is again effected with the aid of a sawtooth voltage. At the beginning of a half-cycle, that is to say at the zero crossing, the generator of the sawtooth voltage is switched on; after the half-cycle at the next zero crossing, the sawtooth voltage is momentarily connected to a capacitor, e.g. by means of a field-effect transistor, and stored in the said capacitor. The half-period duration T/2 is then identical to the voltage value T/2 or analogous to the magnitude of the voltage value. The half-period duration of 1 corresponds to the voltage value u1, that of 2 to that of u2, etc. If pulse amplitude modulation of speech at 8 kHz was effected at the transmitting end, then at the receiving end the voltage u1, u2, u3 must in each case be tapped off at the same frequency and converted into the speech alternating current. In the event of time division multiplex tapping of a plurality of channels, the stored values u1, u2, u3,... must be distributed again with the

same frequency of the time division multiplex tapping. The original information can be produced e.g. by the evaluated code u1, u2,... being formed in a staircase fashion after the channel allocation and this staircase signal being passed via a low-pass filter. Such conversions are known and will not, therefore, be discussed in any more detail.

In the same way as the PDM pulses in Figure 59, PPM pulses can also be decoded. This is illustrated in Figure 60. The distance T2 between the pulses is converted into PAM pulses again by the sawtooth method and stored. The distance T2 then corresponds to the voltage u1, etc.

In the case of the transmission of television signals according to the principle of Figures 36 and 38, the evaluated signals must be distributed synchronously at the receiving end. Synchronizing pulses have to be transmitted in the blanking interval in order that, in accordance with the sampling frequency at the transmitting end, the distribution frequency can be defined at the receiving end. The sum of the longest half-period durations that occur per line must not exceed the time of 54 µs. This is the time provided for a line in the case of a 4:3 picture format. Consequently, the half-period durations must be concomitantly measured in the transmitter. Under certain circumstances, a filling code e.g. comprising the minimum or maximum period durations in a specific sequence must additionally be inserted into the line code. It is also possible, of course, to provide other filling codes. Moreover, the blanking interval can additionally be provided as the filling code as well. Figure 61 illustrates the minimum and maximum half-period durations k and g. Such durations can be transmitted e.g. alternately. Based on this, it is also possible to combine a plurality of channels via one

transmission path. Figure 56 illustrates one such example. The multiplexer Mu combines the channels 1 to n in pulse amplitude terms, this actually being known. These PAM samples are stored in the memory Sp, called up by the PDM and, as already described, fed to the counting element via a control unit St, to which the control pulses Jm are connected. The remaining switching operations are the same as those described e.g. in Figure 36. After the pulse duration modulator PDM, the pulses can also be subjected to further processing directly in accordance with Figure 38. At the receiving end, of course, synchronization and distribution must be effected in accordance with the tapping frequency of the multiplexer.

Figure 57 demonstrates another possibility for multiple utilization of a current path. In order to be able to separate the code alternating currents in frequency terms, control pulses are used which are such that the frequency ranges of the code alternating currents are spaced apart such that entirely satisfactory evaluation is possible, e.g. separation at the receiving location by means of filters. In Figure 57, Z1 is one converter with the control pulses Jm1 and Z2 is the other converter or counting element with the control pulses Jm2. Figure 58 illustrates the frequency of the two channels. T/2I and T/2II are the smallest frequencies of the two channels. As a result of the angular swing f2, the frequency range of the channel T/2I is approximated more closely. In the example, a distance of Ab is also present. This can be chosen such that cost-effective filters can be used.

A few codes which can be used to code and transmit data, television signals in the example, with a frequency are also explained below. Figure 53 illustrates a binary code in which the amplitudes of half-cycles with

the characteristic states of large and small amplitude values are provided as code elements. One bit can then be coded with one half-cycle. Eight bits are provided for the Y signal, in each case 6 bits are provided for the red and blue signals, and 2 bits are provided for the audio (digitized) and other signals. Red and blue are coded alternately, as illustrated e.g. in Figure 51. In the case of 6 Meg tappings for the Y signal, a coding alternating current at 48 MHz would be necessary in this case. Duobinary coding is provided for this purpose in Figure 54. The coding alternating current then has a frequency of 27 MHz. These coding alternating currents can again be transmitted in a frequency-modulated manner; in this case, the frequency band does not become too wide either, as revealed by Figure 64. The transmission reliability becomes even greater in this case. Figure 66 depicts a possible way of digitally transmitting a message in a narrowband manner without modulators. Each code element is assigned a multiplicity of periods of an alternating current at a frequency which are determined by the time Og, that is to say a predetermined number of periods. It is assumed that binary coding is effected. Upon each state change, that is to say 1 to 0 or 0 to 1, the transition takes place continuously designated by Ü in Figure 66. The amplitudes for the zeros have the magnitude Ak and those for the 1s Ag. If identical values occur one after the other, then the amplitude magnitude is not changed; in the case of 5 identical values, a number of periods of Og with the same amplitude would be obtained 5 times. The transition to another characteristic state is classed e.g. as the following characteristic state, that is to say e.g. Ü + O = Og. Figure 65 depicts how the television signals can be digitally arranged serially.

In Figures 53, 54 and 66, the frequency bands

for the transmission of the television signals are very narrow. Under certain circumstances, channels could be accommodated between the individual television channels. The carrier BTz is provided for this purpose in Figure 42. In the case of the coding according to Figure 66, the carrier is simultaneously the modulation signal. In the case of the modulation of the composite video signal with the intermediate frequency carrier 38.9 MHz, in addition to the filter for the generation of the vestigial sidebands, a tuned circuit or series resonant circuit is brought to a frequency such that a curve RR as illustrated in Figure 42 is produced. Such a series resonant circuit is easy to realize. The Nyquist slope should hardly be influenced by this measure.

Claims

1. Method for the digital and/or analog coding of information of one, two or more channels and/or frequency or bandwidth reduction and/or enhancement of the transmission reliability, characterized in that the transmission of information of one, two or a multiplicity of channels is effected with less bandwidth than is made up by the individual channel or the sum of the bandwidths of two or a multiplicity of channels, by the synchronously or quasi-synchronously arranged code elements of the channels to be transmitted being ordered in parallel (Figure 20, S1, S2,..) and thus being combined together to form a code word, and/or in that the digital or analog information items to be coded, if appropriate with the interposition of intermediate stages (e.g. PAM), are converted into PDM pulses, in that, furthermore, means are provided which convert the values of the PDM pulses into the half-period or period durations of half-cycles or periods of a sinusoidal or sine-like alternating current (Figure 35, ER, Figure 36, ER, Figure 38, ER).

2. Method for generating a frequency modulation, characterized in that means are provided which convert an information item or a signal (Figure 30a, Inf) into pulse durations (Figure 30b, 32), in that, furthermore, switching means for measuring the pulse durations, in particular counting switching means (Figure 35, Z), are provided, which simultaneously perform marking of the pulse durations (e.g. Figure 35, Z, A); in this case, the marking circuits are connected in conjunction with pulse duration pulses via gates to an electronic switching means (Figure 35, ER) in such a way that the start and the end of the respective pulse duration pulse code a periodic signal, in particular square-wave pulse;

furthermore, filter means are provided which are such that only sine-like or sinusoidal alternating currents and/or harmonics thereof reach the line (Figure 35, fmo).

3. Method for generating a frequency modulation, characterized in that means are provided which convert an information item or a signal into pulse durations, and in that, furthermore, switching means are provided which convert the duration pulses into an uninterrupted sequence (Pd, Pd, Pd,..) or which convert the pulse duration pulses and the associated intervals (Figure 32, PD1, P, PD2) into, in particular, square-wave pulses (Figures 36, 38), and in that filter means are subsequently provided which are such that they convert these into sinusoidal or sine-like half-cycles or periods to form a coding alternating current.

4. Method according to Claims 1 to 3, characterized in that the pulse duration pulses and intervals or, in the case of storage, pulse duration pulses in an uninterrupted sequence control electronic switching means directly (ER, Figures 36, 38) in such a way that the respective pulse duration or pulse duration interval is converted into a period duration or half-period duration of unipolar or bipolar square-wave pulses, and in that filter means are provided which turn the square-wave pulses into sine-like half-cycles or periods in an uninterrupted sequence of positive and negative half-cycles.

5. Method for evaluating distances e.g. between pulses or half-period or period durations, characterized in that, at the start of the distance marking (Figure 60, 1) or at the zero crossing of the half-period, means for generating a sawtooth voltage are started, and in that, at the end of the distance marking or at the 2nd zero crossing of the half-period (Figure 59), means are connected to the sawtooth voltage which form measurements

thereof or in that means are provided (FET) which store this voltage in a capacitor, in particular.

6. Method according to Claims 1 to 5, characterized in that multiple utilization of current paths is effected by a plurality of information channels being combined in time division multiplex (Figure 56) or by the control pulses for the counting elements obtaining (Figure 57, Jm1, Jm2) a frequency such that their coding alternating currents are not imparted any overlap during the transmission via a current path.

7. Method according to Claim 1, characterized in that, for the coding, a multi-step amplitude code (binary, duobinary, etc.) and/or a phase code or multi-step phase code and/or an analog amplitude and/or phase code is provided, which is provided in particular for the multiple utilization or reduction of the frequency in the case of telex (Figures 18, 19, 20), in the case of television (Figure 21), in the case of teletext, data transmission (Figure 24) and in the case of digital voice transmission (Figure 28).

8. Method for colour television, characterized in that, at the transmitting end, all of the signals are combined in code division multiplex, where the colour, audio and other signals can be assigned as required to a plurality of Y signals in code division multiplex, and in that the receiving end is designed like a superheterodyne receiver, the decoder being arranged downstream of the demodulator (Figure 23, DM), and the decoded signals being distributed in a correctly timed manner by means of the said decoder.

9. Method for the coding of the colour television signals, characterized in that the y signal, red signal y signal, blue signal, Y signal, audio + other signals are tapped off serially in an uninterrupted sequence, in that the PAM values are transferred to the half-period or

period duration of half-cycles or periods of an alternating current, to be precise in the event of amplitude identity, or in that only the sequence Y, r, Y, bl is provided and the audio and other signals are coded by a binary or duobinaryfrequency amplitude code (Figure 55) by each half-cycle or period being assigned an amplitude value which corresponds to the code, in which case the 4 amplitude values (Figure 52) can be assigned to different channels in code division multiplex.

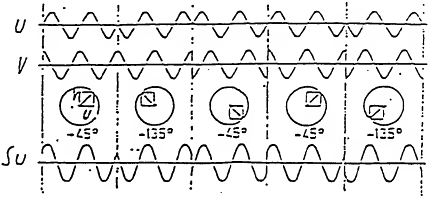
10. Method for the coding of the colour television signals, characterized in that the television signals are only coded with a frequency (Figures 53, 54, 66) by the serially arranged code elements formed by the amplitudes of the half-cycles or periods with the characteristic values of large or small amplitude value or small, medium and large amplitude value being provided for all of the signals, or in that the code is formed from a multiplicity of periods with 2 or 3 characteristic quantities and a continuous transition between the quantities (Figure 66, Ü), this code being provided, as required, for accommodating a channel in the gap between the conventional channels (Figure 42).

11. Method according to Claims 1, 7, 9 and 10, characterized in that the evaluation at the receiving end is effected as far as the decoder as in the case of a superheterodyne receiver.

12. Method according to Claims 1, 7, 8 to 11, characterized in that the television signals are transmitted on the basis of dual QAM, where the y signal is distributed between 2 channels each with 4 bits and these channels are additionally assigned in each case 2 bits for audio and other purposes, the code elements are the half-cycles of an alternating current with the characteristic states of large and small or large, medium and small amplitude values, and the transmission is

effected, as required, on the basis of frequency modulation.

Fig. 2



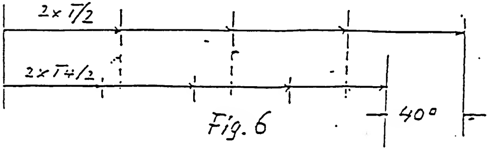


Fig. 6

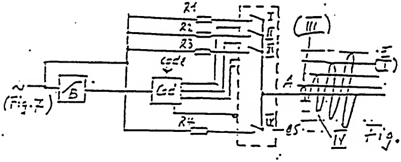


Fig. 9

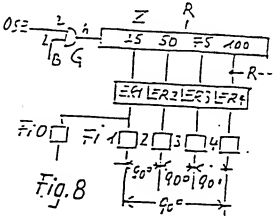


Fig. 8

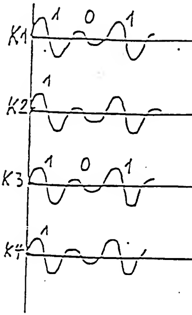


Fig. 10

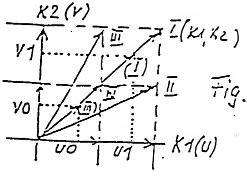


Fig. 11

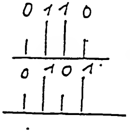
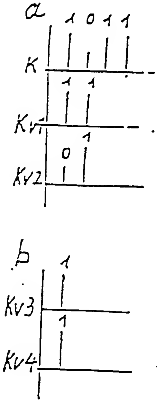


Fig. 13

Fig. 1:



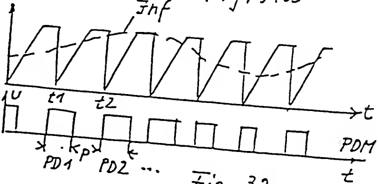


Fig. 31a

Fig. 32a

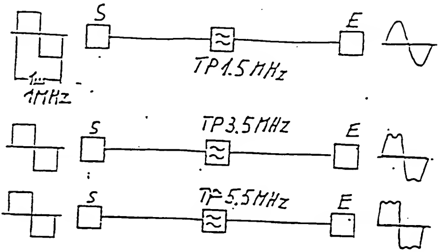
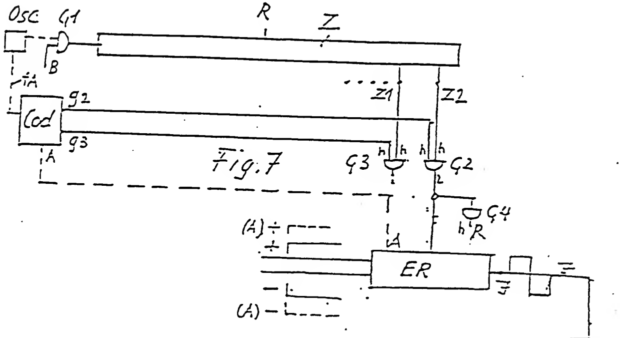


Fig.3



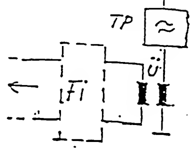


Fig.7

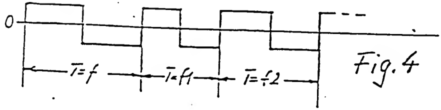


Fig.4

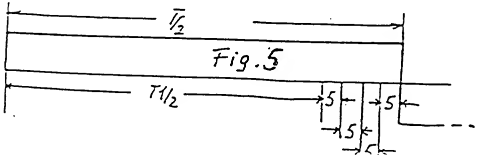


Fig.5

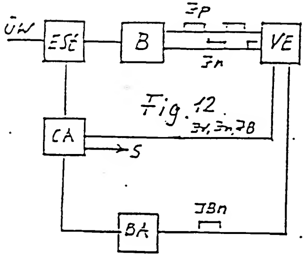


Fig. 12

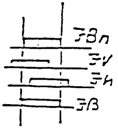


Fig. 12a

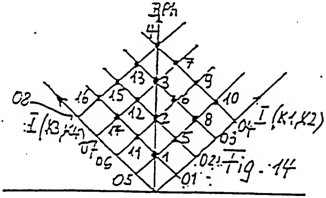


Fig. 14

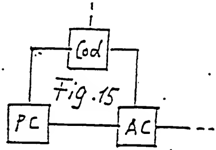


Fig. 15

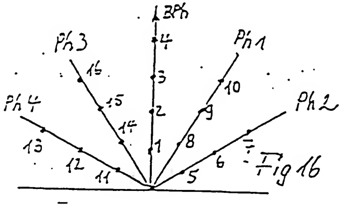


Fig 16

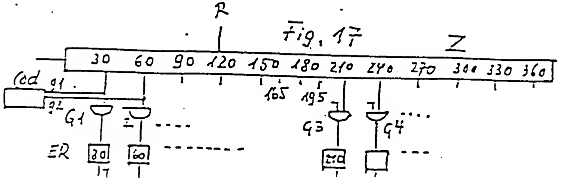


Fig. 17

Fig.18

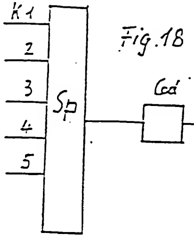


Fig.19

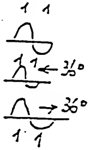
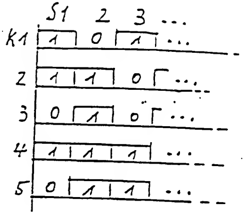


Fig.20



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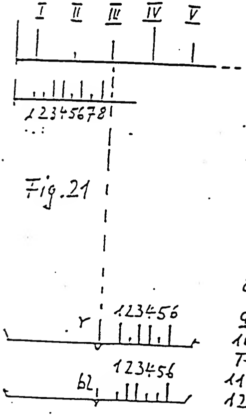
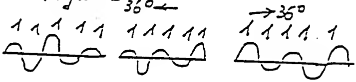


Fig.21

Fig.21b



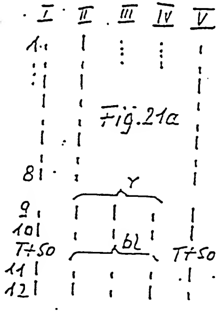


Fig.21a

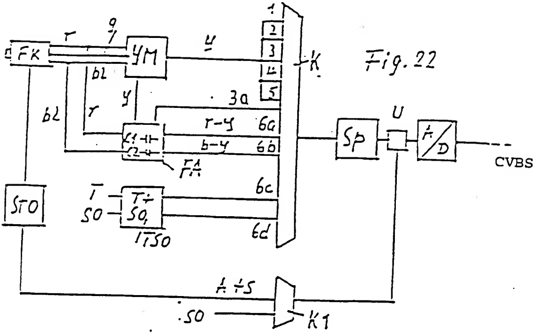


Fig.22

Fig.23

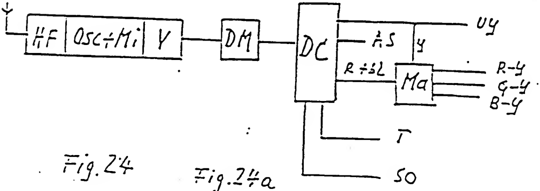


Fig.24

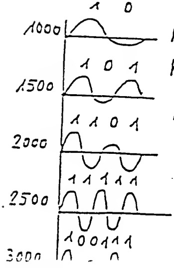


Fig.24a

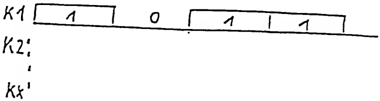


Fig.29a

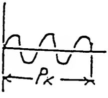
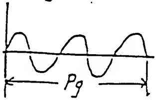


Fig.29b



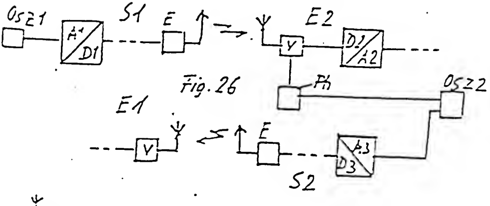


Fig. 26

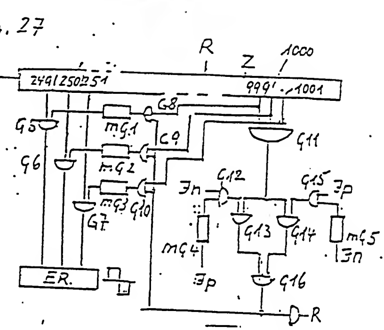


Fig. 27

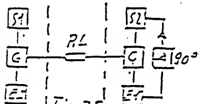


Fig. 25

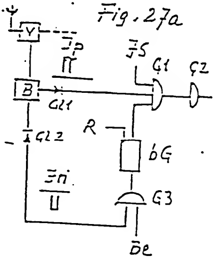
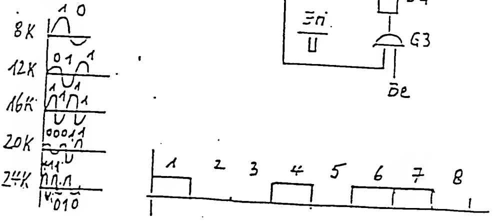


Fig. 27a

Fig. 28



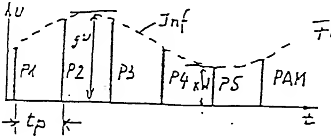


Fig. 30a

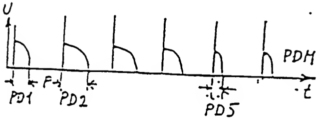


Fig. 30b

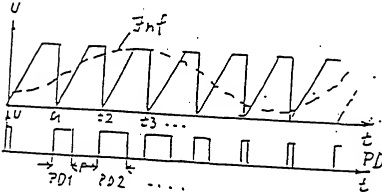


Fig. 31

Fig. 32

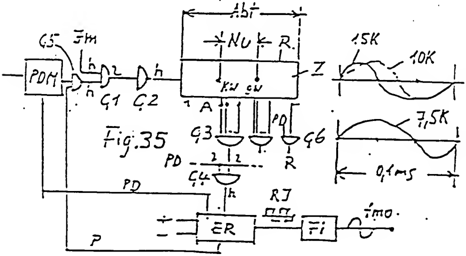


Fig. 35

Fig. 33

Fig. 34

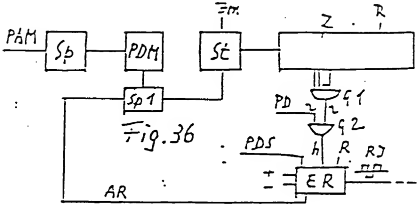


Fig. 36

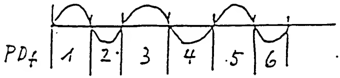


Fig. 37

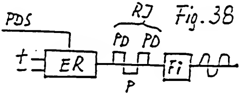


Fig. 38

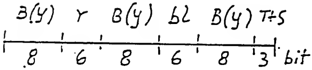


Fig. 65

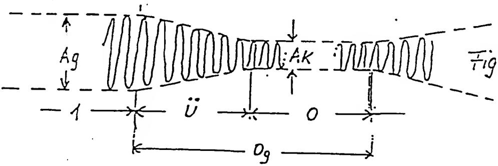


Fig. 66

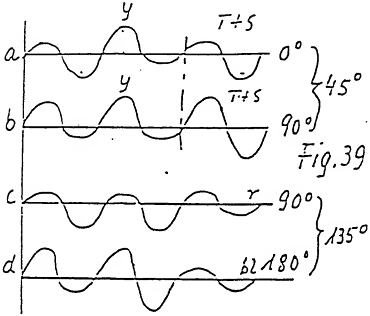
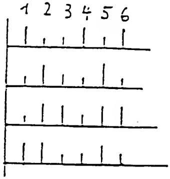


Fig.39

Fig. 40



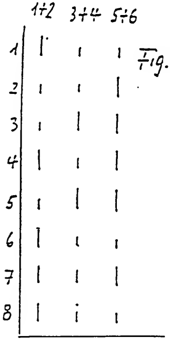


Fig. 41

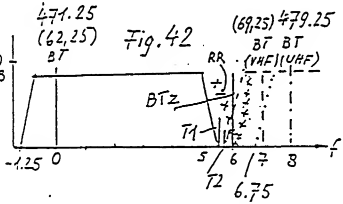


Fig.42

Fig.43

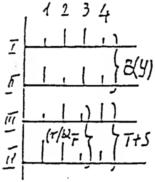


Fig. 44

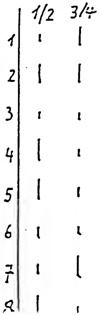


Fig. 45

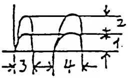


Fig.46

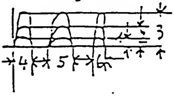


Fig. 47

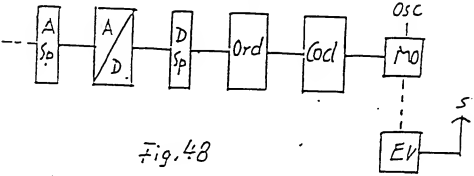
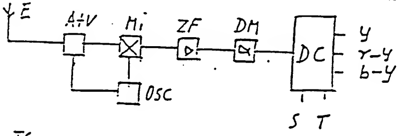


Fig. 48



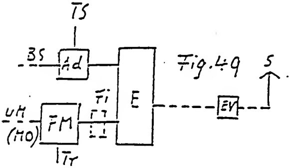


Fig. 49

Fig. 50

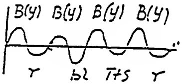


Fig. 51

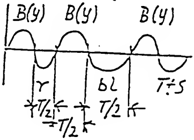
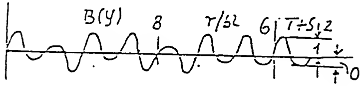


Fig. 53



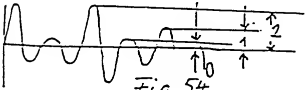


Fig. 54

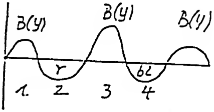


Fig. 52

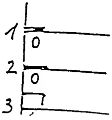


Fig. 55

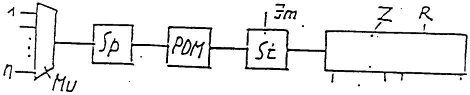
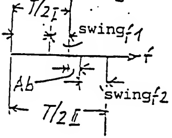


Fig. 56

Fig. 58



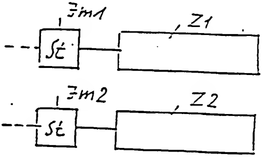


Fig. 57

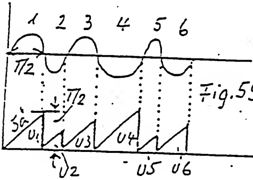


Fig. 59

Fig. 60

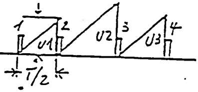


Fig. 61

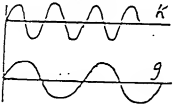


Fig. 62

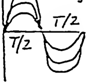
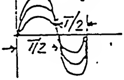


Fig. 63



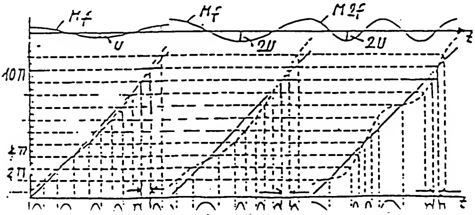


Fig. 64

P10568 11166/P11667

No. 64-5135

SPECIFICATION

1. Title of the Invention

Digital transmission system

2. What is claimed is:

1. A digital transmission system comprising an analog-to-digital converting circuit for converting the signal to be transmitted into a digital code of a desired number of bits, code converting means for taking out a specified number of upper bits of said digital code as a multilevel signal of a relatively small degree of multilevel, and taking out the remaining bits of said digital code as a multilevel signal of a relatively large degree of multilevel, quadrature amplitude modulating means for amplitude-modulating and synthesizing carriers of two axes orthogonal by the multilevel signal of relatively small degree of multilevel and the multilevel signal of relatively large degree of multilevel, and means for transmitting the output of said quadrature amplitude modulating means toward a transmission path.

2. A digital transmission system according to claim 1, wherein error detection correction codes are added to the specified number of upper bits of said digital code and the remaining lower bits, and then they are applied to said code converting means.

3. A digital transmission system according to claim 1,

further comprising means for receiving a signal from said transmission path, means for demodulating the received quadrature amplitude modulation digital signal about two axes, code inverse converting means for taking out the demodulated multilevel signal of relatively small degree of multilevel and multilevel signal of relatively large degree of multilevel as 2-level digital codes, and a digital/analog converting circuit for converting said 2-level digital codes into analog signals.

4. A digital transmission system according to claim 3, wherein a digital signal processing circuit for detecting and correcting the error aid 2-level digital codes occurring during transmission is provided before said digital/analog converting circuit.

3. Detailed Description of the Invention

The present invention relates to a digital transmission system, and more particularly to a transmission system suitable for transmitting digitized voice at high quality.

[Prior Art]

At the present, as exclusive audio broadcast, AM broadcast using medium wave band and FM broadcast using very high frequency band are available. On the other hand, as the compact disc players are distributed widely and digital audio tape recorders are put in practical use today, there is a strong demand for digital broadcast in the field of exclusive audio broadcast.

In this era, the sound broadcasting system by digital

coding has been reported, for example, in "Satellite Broadcast Receiver" in Part 1 of Satellite Broadcast Receiving Technology Investigation Group of Radio Engineering Society, published June 1983, but since reception of satellite broadcast requires a parabolic antenna of about 1 m in diameter, a handy digital audio broadcasting system such as FM broadcast using very high frequency band is requested.

As disclosed in "Satellite Broadcast Receiver", in digital audio, deterioration of transmission signal CN ratio and other transmission error are corrected by using error detection correction code transmitted in superposition, and errors not corrected yet are treated by average interpolation from the preceding and succeeding audio sample values, or the preceding audio sample is held as the previous value. Further, if there are more errors during transmission, it is known to cut off the audio signal output.

[Problems that the Invention Is to Solve]

In the prior art, since no consideration is given to distribution of transmission information into upper bits and lower bits after digital coding, if the CN of the transmission path becomes small and the error rate of the transmission digital code increases, unusual sound is generated or reproduction sound is cut off, and the content of transmission information could not be understood.

The invention is devised to solve the above problems in the quadrature amplitude modulation digital transmission system for amplitude-modulating two orthogonal carriers by two

sets of digital codes. That is, the present inventors promoted studies about these problems, and discovered that occurrence of error rate due to lowering of transmission CN ratio is higher as the degree of multilevel is larger, and that the relatively important bits, that is, upper bits must be lowered in occurrence of error rate as compared with relatively less important bits, that is, lower bits, and attempted to solve the problems.

It is hence an object of the invention, in the quadrature amplitude modulation digital transmission system, to reproduce the information in high quality state if the transmission CN ratio is large and the error rate of transmission digital code is low, and to minimize occurrence of error in the digital code portion having a serious effect on reproduction information if the transmission CN ratio is lowered and the error rate of transmission digital code is entirely increased, thereby reproducing at such an extent as to understand the content of the transmission information.

[Means for Solving the Problems]

To achieve the object, in the quadrature amplitude modulation digital transmission system of the invention, carriers of two orthogonal axes are different in the degree of multilevel of multilevel signal to be modulated, and higher bits of the code converted from analog to digital are assigned to the specified number of bits as the multilevel signal of the axis of smaller degree of multilevel, while the remaining lower bits are assigned as the multilevel signal of the axis of larger

degree of multilevel.

[Operation of the Invention]

When the transmission CN ratio of transmission signal becomes smaller, the error rate of the lower bits transmitted at larger degree of multilevel is higher, but the error rate is lower in the upper bits transmitted at smaller degree of multilevel.

Since the error rate of upper bits is lower, large error of amplitude is rare in analog signal, and extremely unusual sound is hardly generated, and it is not required to cut off reproduction sound, and reproduction sound of such a quality as to understand the content of transmission information can be obtained.

[Embodiment]

As an embodiment of the invention, an example of three-bit transmission is explained below, in which the number of transmission bits of the quadrature amplitude modulation (hereinafter called QAM) is 4 bits, and the Q-axis of 16QAM is 2-level. Fig. 1 shows an example of a receiving and reproducing apparatus of the invention, in which reference numeral 1 is an antenna, 2 is a channel selection circuit, 3 is a first synchronous detection circuit, 4 is a second synchronous detection circuit, 5 is a carrier regenerating circuit, 6 is a phase shifter, 7, 8 are LPFs (low pass filters), 9 is a first discriminating circuit, 10 is a second discriminating circuit (4-level-2-level converting circuit), 12 is a first receiving side digital signal processing circuit, 13 is a second receiving

side digital signal processing circuit, 14 is a digital/analog converting circuit (hereinafter called DAC), and 15 is an audio output. Fig. 2 shows an example of a transmitting side transmission signal generating apparatus of the invention, in which reference numeral 21 is an audio input, 22 is an analog-to-digital converting circuit (hereinafter called ADC), 23 is a first transmitting side digital signal processing circuit, 24 is a second transmitting side digital signal processing circuit, 25 is a 2-level-4-level converting circuit, 26, 27 are LPFs, 28 is a carrier generating circuit, 29 is a phase shifter, 30 is a first modulating circuit, 31 is a second modulating circuit, 32 is an adder, 33 is an amplifier, and 34 is an antenna. Fig. 3 shows a code layout example of transmission signal of the invention, and Fig. 4 is a bit distribution example of transmission signal of the invention.

The operation is explained first from the receiving side.

The transmitted wave is received in the antenna 1 in Fig. 1, and the broadcasting station is selected in the channel selection circuit 2. The intermediate frequency signal after channel selection is synchronously detected in the quadrature relation by the first synchronous detecting circuit 3 and second synchronous detecting circuit 4, by the output of the carrier regenerating circuit 5 and output of the phase shifter 6, and undesired signals are removed by the LPF 7 and 8. As the output, the Q-axis has an eye pattern of 2-level value, and the I-axis has one of 4-level value. From the eye patterns, 2-level digital codes are obtained by the output of the clock

regenerating circuit 11 and the first discriminating circuit 9 and second discriminating circuit 10. Then, in the first digital signal processing circuit 12 and second digital signal processing circuit 13, detection and correction of error occurring during transmission, de-interleaving, and digital signal processing for demodulating digital transmission are executed, and the code is converted into an analog signal in the DAC 14, and an audio output 15 is obtained.

Referring next to Fig. 2, the transmitting side operation is explained. Fig. 2 is a block diagram of an apparatus for generating a transmission signal for reproducing in this receiving and reproducing apparatus. The analog signal from the audio input 21 is converted into a 2-level digital code in the ADC 22, codes for detecting and correcting errors occurring during transmission are added by the first digital signal processing circuit 23 and second digital signal processing circuit 24, and interleaving or other process is done to avoid burst error. Then, on the I-axis, the 2-level output of the second digital processing circuit 24 is applied into the 2-level-4-level converting circuit 25 to be converted into a 4-level value, the undesired band is removed through the LPF 27, and the output of the carrier generating circuit 28 is modulated in the second modulating circuit 31 by using the signal shifted in phase by 90 ° through the shaft shifter 29. On the other hand, on the Q-axis, the 2-level output of the first digital processing circuit 23 is applied to the LPF 26 to remove undesired band, and is modulated in the first modulating circuit

30 by using the output of the carrier generating circuit 28. In this embodiment, the degree of multilevel of Q-axis remains at 2-level, the 2-level-multilevel converting circuit as on the I-axis is omitted. The outputs of the converting circuits 30, 31 are added in the adder 32, and amplifier in the amplifier 33, and transmitted as radio wave from the antenna 34.

Fig. 3 shows the code layout of QAM signal by modulating the I-axis by 4-level and Q-axis buy 2-level. The axis of abscissas in Fig. 3 is the Q-axis, expressed by 2-level of 0 and 1, and the I-axis is 4-level of 00, 01, 10, 11, so that three-bit data can be simultaneously transmitted in a same time slot. This is shown in Fig. 3 in the sequence of (Q, I_1, I_2). There is a difference of three times between the inter-code distance on the Q-axis and the inter-code distance on the I-axis, and the transmission signal CN ratio at which the bit error rate is the same may be smaller by 10 dB on the Q-axis. In other words, in the case of a signal transmitted at a certain CN ratio, the error rate is smaller on the Q-axis.

Further, as shown in Fig. 4, assuming the audio signal to be transmitted is quantized at N bits per 1 sample, for example, 12 bits, the data is supposed to be D_1, D_2, D_3, ..., D_{12} sequentially from the highest bit (MSB), and in the higher M bits, the error detection and correction code is, for example, E_1 for three bits D_1 to D_3, E_2 for D_4 to D_6, and E_3 for D_7 to D_9. In this case, in the time of time slots T_1 to T_5, by distributing D_1 to D_4 and E_1 to Q, and D_5 to D_{12}, E_2 and E_3 to I_1 and I_2, in the case of deterioration of CN ratio of transmission signal,

since the higher bit side is assigned to the Q-axis, the error rate is low, while the lower bit side is assigned to the I-axis, and the error rate is high. As a result, if the CN ratio of transmission signal is extremely poor and the lower bits are nearly completely errors, the error rate of the upper bits is small, and the audio signal be reproduced to a certain extent.

The error detection and correction codes E_1 to E_3 are indicated as parity for 1 sample, but an error detection and correction code of several bits may be assigned by collecting upper three bits of several samples.

As described herein, according to the embodiment, if the transmission CN ratio is large and the error rate of transmission digital code is large, reproduction sound of 12 bits is obtained, and if the CN ratio is small and poor, the error is small in the upper three or four bits, so that the reproduction sound is obtained to such an extent as to be understood as voice.

In the case of three-bit transmission, the required transmission band width is calculated. Supposing the number of quantized bits to be 12 bits, the sampling frequency to be 32 kHz, the sound channels to be two (stereo), and the error correction code superposition to be 30%,

$$12 \text{ bits} \times 32\text{K} / \text{S} \times 2\text{ch} \times 1.3 = 998.4 \text{ kbps}$$

and 998.4 kbps (kbits/sec) is obtained, and by simultaneous three-bit transmission, it is 332.8 kbps, which can be transmitted in a band width of 332.8 kHz. This band width is similar to that of the existing FM broadcast, and it can be

transmitted in the very high frequency band.

On the other hand, the carrier regenerating circuit 5 is important for obtaining the regenerative orthogonal axes, and on the basis of only the 4-level case of data (0, 0, 0), (0, 1, 1), (1, 0, 0),and (1, 1, 1), a method of negative feedback so that the amplitude may be equal on the I-axis and Q-axis may be considered. This circuit is, in the case of 16QAM, explained in the reference carrier regenerating circuit shown in pp. 134-135 of "Digital Microwave Communications", published by Project Center, May, 1984.

The audio signal is explained so far, but same effects are obtained in video signal and other data in which upper bits present important information.

Herein, four bits of 16QAM are transmitted in three bits in eight states, but same effects are obtained in other QAM such as transmission of six bits of 64QAM in five bits in 32 states by 8-level on the I-axis and 4-level on the Q-axis. In this case, the transmitting side requires 2-level-8-level converting circuit on the I-axis and 2-level-4-level converting circuit on the Q-axis. Similar reverse converting circuits are also required at the receiving side.

[Effects of the Invention]

As the embodiment is described herein, according to the quadrature amplitude modulation digital transmission system of the invention, carriers of two orthogonal axes are different in the degree of multilevel of multilevel signal to be modulated, and higher bits of the code converted from analog to digital

are assigned as the multilevel signal of the axis of smaller degree of multilevel, while the remaining lower bits of the code converted from analog to digital are assigned as the multilevel signal of the axis of larger degree of multilevel, and therefore reproduction at high quality is possible when the transmission CN ratio is large and the transmission condition is favorable, and in the poor condition of lower transmission CN ratio, increase of error rate is suppressed in the upper bits as compared with lower bits, so that it is possible to reproduce to such an extent as the content of the transmission information can be understood, and many other excellent effects are brought about.

4. Brief Description of the Drawings

Fig. 1 is a block diagram of an embodiment of receiving and reproducing apparatus according to the invention, Fig. 2 is a block diagram of an embodiment of transmitting side transmission signal generating apparatus of the invention, Fig. 3 is a diagram showing an example of code layout of transmission signal used in the invention, and Fig. 4 is a diagram showing an example of bit layout of transmission signal used in the invention.

3, 4 Synchronous detector

9, 10 Multilevel code discriminating circuit (4-level-2-level converting circuit)

11 Clock regenerating circuit

12, 13, 23, 24 Digital signal processing circuit

14 Digital-to-analog converting circuit

21 Audio input terminal

22 Analog-to-digital converting circuit

25 2-level-4-level converting circuit

30, 31 Quadrature modulation circuit

32 Adder

Attorney: Katsuo Ogawa, patent attorney

Fig. 1

2 Channel selector

5 Carrier regenerator

9 Discriminating

10 Discriminating 4-level-2-level conversion

11 Clock regenerator

12 Digital signal processor

13 Digital signal processor

15 Output

Q-axis

I-axis

Fig. 2

23 Digital signal processor

24 Digital signal processor

25 2-level-4-level converter

33 Amplifier

Fig. 3

Q-axis

I-axis

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㊹発明の名称　ディジタル伝送方式

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㉒出　願　昭62(1987)6月29日

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明　細　書

1. 発明の名称

ディジタル伝送方式

2. 特許請求の範囲

1. 伝送すべき信号を所要ビット数のディジタル符号に変換するアナログディジタル変換回路と、該ディジタル符号の上位所定数のビットを比較的多値化の程度の少ない多値信号として取出すと共に、該ディジタル符号の残りのビット数を比較的多値化の程度の大きい多値信号として取出す符号変換手段と、前記比較的多値化の程度の少ない多値信号及び比較的多値化の程度の大きい多値信号により直交した2軸の搬送波をそれぞれ振幅変調して合成する直交振幅変調手段と該直交振幅変調手段の出力を伝送路に向けて送出する手段とを備えたディジタル伝送方式。

2. 前記ディジタル符号の上位所定数のビット及び残りの下位ビットにそれぞれ誤り検出訂正符号を付加してから前記符号変換手段に印加するようにした特許請求の範囲第1項記載のディジタル伝送方式。

3. 前記伝送路からの信号を受信する手段、受信された直交振幅変調ディジタル信号を2軸について復調する手段と、復調された比較的多値化の程度の少ない多値信号及び比較的多値化の程度の大きい多値信号をそれぞれ2値ディジタル符号として取出す符号逆変換手段と、該2値ディジタル符号をアナログ信号に変換するディジタルアナログ変換回路とを備えてなる特許請求の範囲第1項記載のディジタル伝送方式。

4. 前記2値ディジタル符号中の、伝送中に生じた誤りを検出訂正するディジタル信号処理回路を前記ディジタルアナログ変換回路の前に設けてなる特許請求の範囲第3項記載のディジタル伝送方式。

3. 発明の詳細な説明

本発明は、ディジタル伝送方式に係り、特に、ディジタル符号化した音声を高品質で伝送するのに好適な伝送方式に関する。

〔従来の技術〕

現在、オーディオ専用放送として、中波帯を用いたAM放送および超短波帯を用いたFM放送が実施されている。一方、コンパクト・ディスク・プレーヤの普及が進み、ディジタル・オーディオ・テープレコーダが実用化されようとしている今日、このオーディオ専用放送の分野においてもディジタル化の要望が強まってきている。

このような時代において、音声をディジタル符号化して放送する方式については、昭和58年6月発行財団法人電波技術協会編の衛星放送受信技術調査会報告第1部「衛星放送受信機」などで報告されているが、衛星放送受信には直径1m程度のパラボラアンテナを必要とするので超短波帯を用いたFM放送のように手軽に受信できるディジタル・オーディオ放送が望まれる。

また、上記「衛星放送受信機」にも示されているように、ディジタル音声において伝送信号CN比の劣化など伝送中の誤りに対しては重畳して伝送された誤り検出訂正符号を用いて訂正し、訂正しきれないものについては前後の音声サンプル値から平均値補間したり前の音声サンプル値を前値保持したりする。さらに伝送中の誤りが多くなると音声信号出力をしゃ断することが知られている。

〔発明が解決しようとする問題点〕

上記従来技術は、伝送情報をディジタル符号化した後の上位ビットと下位ビットとの誤り率の配分について全く配慮がされていないため、伝送路のCNが小さくなり伝送ディジタル符号の誤り率が多くなると異常音を発生したり再生音を遮断したりするので、伝送情報内容を理解できない問題があった。

本発明は、直交する2つの搬送波を2組のディジタル符号で振幅変調する直交振幅変調ディジタル伝送方式において、上記の問題を解決するためになされたものである。すなわち、本発明者は、この問題について研究を進めた結果、伝送CN比の低下による誤り率の発生は、多値化の程度の大きい程多くなることと、並びに、比較的重要なビットすなわち上位のビットは、比較的重要でないビットすなわち下位ビットに比べて誤り率の発生をより少くする必要があることに着目して、その解決を図ったものである。

従って、本発明の目的は、直交振幅変調ディジタル伝送方式において、伝送CN比が大きく伝送ディジタル符号の誤り率が少ない場合には高品質な状態でもとの情報を再生し、伝送CN比が低下して伝送ディジタル符号の誤り率が全体として多くなった場合でも、再生情報に重要な影響を与えるディジタル符号部分の誤りの発生を極力抑えるようにして、伝送情報内容が理解できる程度の再生を可能とするものである。

〔問題点を解決するための手段〕

上記目的を達成するため、本発明の直交振幅変調ディジタル伝送方式においては、直交した2軸の搬送波をそれぞれ変調する多値信号の多値化の程度を異ならしめ、多値化の程度の少ない軸の多値信号として所要ビット数にアナログディジタル変換された符号の上位ビットを割り当て、多値化の程度の多い軸の多値信号として上記符号の残りの下位ビットを割り当てるように構成する。

〔作用〕

伝送信号の伝送CN比が小さくなると、多い多値化で伝送される下位ビットの誤り率が多くなるが、少ない多値化で伝送される上位ビットの誤り率は少ない。

上位ビットの誤り率が少ないため、アナログ信号で振幅を大きく誤ることが少なく、ひどい異常音を発生することが少ないため、再生音を遮断する必要もなく、伝送情報の内容を理解できる再生音を得られる。

〔実施例〕

以下、本発明の一実施例として直交振幅変調(以下QAMと略す)の伝送ビット数を4ビットの16QAMのQ軸を2値化にした3ビット伝送を例にとり説明する。第1図に本発明の受信再生装置の一実施例であり、1はアンテナ、2は選局回路、3は第1の同期検波回路、4は第2の同期検波回路、5は搬送波再生回路、6は移相器、7,8はLPF(低域通過フィルタ)、9は第1の識別回路、10は第2の識別回路(4値−2値変換回

路）、12は第1の受[illegible]ディジタル信号処理回路、13は第2の受信側ディジタル信号処理回路、14はディジタル・アナログ変換回路（以下DACと略す）、15は音声出力である。第2図は本発明の送信側の伝送信号発生装置の一実施例であり、21は音声入力、22はアナログ・ディジタル変換回路（以下ADCと略す）、23は第1の送信側ディジタル信号処理回路、24は第2の送信側ディジタル信号処理回路、25は2値－4値変換回路、26，27はLPF、28は搬送波発生回路、29は移相器、30は第1の変調回路、31は第2の変調回路、32は加算回路、33は増幅器、34はアンテナである。第3図は本発明の伝送信号の符号配置例、第4図は本発明の伝送信号のビット配分例を示す。

都合により、まず、受信側から動作を説明する。

伝送された電波を第1図のアンテナ1で受け、選局回路2で放送局を選局する。選局された後の中間周波信号を搬送波再生回路5の出力と移相器6の出力により第1の同期検波回路3と第2の同期検波回路4とで[illegible]の直交関係で同期検波し、LPF7および8で不要信号を除去する。その出力として、Q軸は2値、I軸は4値のアイパターンを得ている。そのアイパターンからクロック再生回路11の出力と第1の識別回路9および第2の識別回路10により2値のディジタル符号を得る。その後、第1のディジタル信号処理回路12および第2のディジタル信号処理回路13で伝送中に生じた誤りの検出訂正やデインタリーブなどディジタル伝送を復調するディジタル信号処理を行い、DAC14でアナログ信号にして音声出力15を得る。

次に、第2図により、送信側の動作を説明する。第2図は、以上の受信再生装置で再生するための伝送信号を発生する装置のブロック図である。音声入力21からのアナログ信号をADC22で2値のディジタル符号化し、第1のディジタル信号処理回路23および第2のディジタル信号処理回路24により、伝送中に生じる誤りを検出訂正するための符号を追加し、また、バースト誤りをさけるためインターリーブなどをほどこす。その後、I軸では第2のディジタル処理回路24の2値出力は4値化するため、2値－4値変換回路25に印加され、LPF27を通って不要帯域が除去され、搬送波発生回路28の出力を移相器29を介して90°移相した信号を用いて第2の変調回路31で変調される。一方、Q軸では第1のディジタル処理回路23の2値出力が直接LPF26に印加されて不要帯域が除去され、搬送波発生回路28の出力を用いて第1の変調回路30で変調される。なお、この実施例では、Q軸の多値化の程度を2値そのままとしたため、I軸のような2値－多値変換回路は省略されている。それらの変調回路30，31の出力を加算器32で加算し、増幅回路33で増幅してアンテナ34から電波として伝送する。

このようにI軸を4値、Q軸を2値で変調したQAM信号の符号配置を第3図に示す。第3図の横軸がQ軸であり0と1の2値、I軸は00，01，10，11の4値となり3ビットのデータを同時に同一タイムスロットで伝送できる。これを（Q，I_1，I_2）の順で第3図に示す。ここでQ軸の符号間距離とI軸の符号間距離には3倍の差があり、ビット誤り率が同一となる伝送信号CN比はQ軸の方が10dB少なくて良い。逆に言えばあるCN比で伝送された信号の場合Q軸の方が誤り率が少ないことになる。

今、第4図に示すように、伝送する音声信号を1サンプル当りNビット例えば12ビットで量子化したと仮定し、そのデータを上位ビット（MSB）から順に$D_1, D_2, D_3, \cdots D_{12}$とし、上位Nビット例えば3ビット$D_1 \sim D_3$について$E_1$，$D_4 \sim D_6$について$E_2$，$D_7 \sim D_9$について$E_3$の誤り検出訂正符号とする。このときタイムスロット$T_1 \sim T_5$の時間において、Qに$D_1 \sim D_4$と$E_1$を、$I_1$，$I_2$に$D_5 \sim D_{12}$と$E_2$と$E_3$を配分することにより、伝送信号のCN比が劣化した場合上位ビット側はQ軸に割り当てられているため誤り率は少なく、下位ビット側はI軸に割り当てられているため誤り率が多くなる、その結果、極端に伝送信号のCN比が劣化して、下位ビットのほとんどが誤りとなったとしても、上位ビ

ットの誤り率が少なく、ある程度の音声信号を再生できる。

なお、E_1～E_3の誤り検出訂正符号を1サンプルについてのパリティのように示したが、数サンプルの上位3ビットをまとめて数ビットの誤り検出訂正符号をつけても良い。

以上説明したように、本実施例によれば、伝送CN比が大きく伝送ディジタル符号の誤り率が大きい場合には、12ビットの再生音が得られ、CN比が小さくなり悪くなった場合でも上位3～4ビットは誤り少なく得られるので、音声として理解できる程度の再生音が得られる効果がある。

ここで、3ビットで伝送した場合の伝送必要帯域幅を計算する。量子化ビット数12ビット、サンプリング周波数32KHz、音声2チャネル(ステレオ)、誤り訂正符号重畳分を30％とすると、

12bit×32K/S×2ch×1.3=998.4Kbps

998.4Kbps(Kビット／秒)となり同時3ビット伝送するので332.8Kbpsとなり332.8KHzの帯域幅で伝送可能となる。この帯域幅は現行FM放送と同程度で、超短波帯で伝送可能である。

一方、搬送波再生回路5は再生直交軸を得るために重要であり、データ(0,0,0),(0,1,1),(1,0,0)および(1,1,1)の4値の場合のみを基準としてI軸，Q軸への振幅が同一となるように負帰還する方法が考えられる。この回路は、16QAMの場合には、昭和59年5月に株式会社企画センター発行の「ディジタルマイクロ波通信」のpp134～135に示した基準搬送波再生回路に説明されている。

以上、音声信号で説明したが、画像信号など上位ビットが重要情報を有するものについても同様な効果がある。

また、今までの説明では16QAMの4ビットを8状態の3ビットにして伝送したが、64QAMの6ビットのI軸を8値としQ軸を4値とした32状態の5ビットにした伝送など他のQAMでも同様な効果が得られる。なお、この場合には、送信側で、I軸に2値－8値変換回路、Q軸に2値－4値変換回路が必要になる。受信側でも同様な逆変換回路が必要である。

〔発明の効果〕

以上実施例により詳述したように、本発明の直交振幅変調ディジタル伝送方式によれば、直交2軸の搬送波をそれぞれ変調する多値信号の多値化の程度を異ならしめ、多値化の程度の少ない軸の多値信号としてA/D変換された符号の上位ビットを配分し、多値化の程度の多い軸の多値信号としてA/D変換された符号の残りの下位ビットを配分したので、伝送CN比が大きくて良質な伝送条件のときには高品質な再生ができ、伝送CN比が低下した悪条件においても、下位ビットに比べて上位ビットの誤り率の増加を極力抑えることができ、その結果、伝送情報内容が理解できる程度の再生を可能とする等、優れた効果を奏するものである。

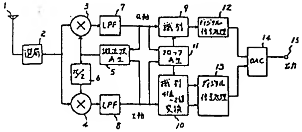
4. 図面の簡単な説明

第1図は本発明に用いる受信再生装置の一実施例のブロック図、第2図は本発明の送信側の伝送信号発生装置の一実施例のブロック図、第3図は本発明に用いる伝送信号の符号配置の一例を示す図、第4図は本発明に用いる伝送信号のビット配分の一例を示す図である。

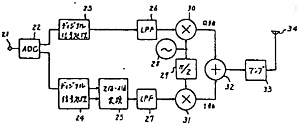
3，4…同期検波器
9，10…多値符号識別回路(4値－2値変換回路)
11…クロック再生回路
12，13，23，24…ディジタル信号処理回路
14…ディジタル・アナログ変換回路
21…音声入力端子
22…アナログ・ディジタル変換回路
25…2値－4値変換回路
30，31…直交変調回路
32…加算回路。

代理人 弁理士 小川勝男

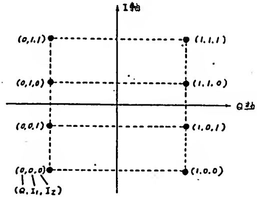
第1図



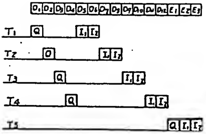
第2図



第3図



第4図



P10[illegible]8/P 11166/P 11667

No. 63-28145

SPECIFICATION

1. Title of the Invention

Wireless communication system

2. What is claimed is:

1. A wireless communication system, being a system of presenting a plurality of services differing in the required transmission quality by wireless communication,

wherein each service is presented by wireless communication, by a same transmitter and a same transmission power, and

the service signal is provided with the transmission characteristic improving treatment for obtaining a greater improving effect when the requirement is stricter depending on the required transmission quality of the service.

3. Detailed Description of the Invention

[Industrial Field of Utilization]

The present invention relates to a wireless communication system for presenting a plurality of services, and more particularly to a wireless communication system suited to a mobile communication system.

[Prior Art]

In mobile communication, when presenting a plurality of services (for example, sound, facsimile, and data

communication), it is supposed that the required transmission quality (such as bit error rate) differs individually.

In mobile communication, usually, a radio base station connected to a fixed communication network is installed in the center of service area, mobile stations moving freely in the service area are connected to the fixed communication network through the radio base station. The communication range of mobile stations (called zone radius) is determined by the transmission required in communication and transmission electric power between the base station and mobile stations.

Generally, in facsimile or data communication, stricter transmission quality is required than in voice communication, and therefore in the system having set the transmission electric power for voice communication, if desired to receive the service of facsimile or data communication by using the same transmitter and receiver, the user cannot receive the service of facsimile or data communication except for the central region of the service area. Accordingly, to realize facsimile or data communication in satisfactory quality in the entire region of voice communication, the transmission electric power must be increased at the time of facsimile or data communication.

It is relatively complicated to control the transmission electric power in every service, and when the transmission electric power is increased, the distance of the wireless communication system using the same frequency must be set apart, and the frequency utilization efficiency is poor. In particular, in mobile communication, the service area using a

same frequency must be extended in distance, and the effective use of frequency becomes poor.

It is hence an object of the invention to present a wireless communication system capable of presenting a plurality of services differing in transmission quality, by a same transmitter and same transmission electric power in a same area.

[Means for Solving the Problems.]

According to the invention, service signals differing in the required transmission quality are transmitted by a same transmitter and same transmission electric power, and the service signals are treated by transmission characteristic improvement differing depending on the required transmission quality, and in this case, the stricter the required transmission quality, the greater is the obtained improvement effect.

Thus, for all services, for example, communication can be in the same zone radius and same transmission electric power

[Embodiment]

Fig. 1 shows an example of mobile communication system for explaining an embodiment of the invention. A voice signal input terminal 1, a facsimile signal input terminal 2, and a data communication input terminal 3 are connected to a switch 5 through a signal processing circuit 4 for improvement of transmission characteristic. In this embodiment, the transmission characteristic improvement technology is realized by error correction coding and time diversity, and the signal input terminals 1, 2, 3 are respectively connected to

error correction coding circuits 4a, 4b and 5c in the signal processing circuit 4, output sides of the error correction coding circuits 4a, 4b and 4c are respectively connected to time diversity circuits 4d, 4e and 4f, and these time diversity circuits 4d, 4e and 4f are connected to a transmitter 6 through the switch 5. The transmission signal of the transmitter 6 is transmitted as radio wave from a transmission antenna 7.

This radio wave is received in a reception antenna 8, and is supplied into a receiver 9. The output side of the receiver 9 is changed over and connected to any one of the circuits corresponding to voice signal, facsimile signal and data signal in a signal processing circuit 11 for improvement of transmission characteristic through a switch 10. The signal processing circuit 11 includes a voice signal output terminal 12, a facsimile signal output terminal 13, and a data signal output terminal 14.

A coded voice signal is fed into the voice signal input terminal 1. The coded voice signal is provided with a check bit by the error correction coding circuit 4a, and the time diversity circuit 4d sends out the same signal plural times at intervals (as for operation of time diversity, see Japanese Laid-open Patent No. 56-191814). The facsimile signal and data signal, similarly, pass through the error correction coding circuits 4b, 4c and time diversity circuits 4e, 4f, and are fed into the switch 5. The switch 5 selects any one of voice signal, facsimile signal and data signal, and supplies it into the transmitter 6, and this signal is modulated in carrier in the

transmitter 6, and is transmitted to the transmission antenna 7.

The transmission signal is received in the reception antenna 8, and is demodulated and decoded into a base band signal in the receiver 9, and is put into the signal processing circuit 11. The signal processing circuit 11 is a circuit for processing reversely as in the signal processing circuit 4, being provided individually for voice signal, facsimile signal and data signal, and each demodulated and decoded signal is processed by time diversity and error correction coding, and the voice signal is issued from the voice signal output terminal 12, the facsimile signal from the facsimile signal output terminal 13, and the data signal from the data signal output terminal 14.

In this case, according to the invention, the voice signal, facsimile signal, and data signal are processed by correction coding at different correction capacity and time diversity of different number of branches, individually, that is, the higher the required transmission quality, the higher is raised the correction capacity of error correction coding and the larger is the number of branches of time diversity. For example, the correction capacity of error correction coding is higher and the number of branches of time diversity is larger in the facsimile signal than in voice signal.

Thus, plural services of different transmission quality requirements can be presented by same transmission electric power and in same zone radius.

Depending on the requirement of transmission quality, meanwhile, only the correction capacity of error correction coding or only the number of branches of time diversity may be varied.

[Effects of the Invention]

The effects of the invention are described below while referring to specific examples. Supposing the voice signal to be an analog signal of 3 kHz coded according to APC-AB (adaptive prediction-adaptive bit assignment), the facsimile signal to be a signal of 4.8 kb/s of G3, and the data signal to be a signal of 2.4 kb/s, their required transmission quality is respectively assumed to be 10^{-2}, 10^{-4}, and 10^{-5}. Using two-branch spatial diversity (2SD) as fading measure, in the case of voice signal, at the transmission electric power of 15 W/3 W in the base station/mobile station, the frequency assignment for service area of zone radius of 3 km in 1.5 GHz band is realized by repeating nine sets of frequency. In the case of facsimile signal, however, at the same transmission electric power, the frequency assignment for service area of zone radius of 1.4 km realized by repeating 36 sets of frequency.

As shown in Fig. 2, the voice signal from the input terminal 1 is coded in an APC-AB coding circuit 15, and is also coded by bit sort error correction (BSFEC), and the coded voice signal is sent out into the switch 5 at 16 kb/s. The facsimile signal is coded in the error correction coding circuit 4b, and fed into the time diversity circuit 4e to undergo time diversity of two branches (2TD), and is supplied into the switch 5 at 16 kb/s.

That is, since the time diversity has two branches, 8 kb/s is issued from one branch, and its 3 (8-4.8) kb/s is used in error correction bit. The data signal from the terminal 3 is coded in the error correction coding circuit 4c, and is fed into the time diversity circuit 4f to undergo time diversity of four branches (4TD), and is supplied into the switch 5 at 16 kb/s. The signal is modulated by GMSK (BbT = 0.25) and transmitted in a transmitter-receiver 21. That is, the transmission speed in the wireless section is 16 kb/s. The signal is received by a two-branch spatial diversity antenna 22, and demodulated in the transmitter-receiver 21 by frequency detection two-bit integral detection system, and is decoded by supplying into any one of the coding circuit 15, time diversity circuits 4e, 4f, through the switch 5.

Fig. 3 shows measured results of experiments of average bit error rate with respect to the reception CNR (central value) in the case of using only two-branch spatial diversity in the presence of Raleigh fading (2SD), in the case of using two-branch spatial diversity, two-branch time diversity and error correction coding (2SD-2TD-FEC), and in the case of using two-branch spatial diversity, four branch time diversity and error correction coding (2SD-4TD-FEC).

As known from Fig. 3, at the reception CNR of near 10 dB, the voice signal has an average bit error rate of 10^{-2} by 2SD, the facsimile signal has an average bit error rate of 10^{-4} by 2SD-2Td-FEC, and the data signal has an average bit error rate of 10^{-5} by 2SD-4TD-FEC. That is, when the voice signal,

facsimile signal, and data signal are treated by transmission characteristic improvement as shown in Fig. 2 individually, the required transmission quality is obtained at the same transmission electric power. When applied to the mobile wireless communication, at the zone radius of 3 km, the frequency assignment for service area can be realized by repeating nine sets of frequency, and not only the voice signal, but also the service of facsimile signal and data signal can be presented.

As described herein, according to the invention, in the area capable of transmitting, for example, voice by the same transmitter and same transmission electric power, the service of facsimile or data communication is realized, and the user can enjoy a plurality of services without being conscious of difference in service. This invention can be applied not only in mobile communications but also in general wireless communications.

4. Brief Description of the Drawings

Fig. 1 is a block diagram showing a wireless communication system according to the invention, Fig. 2 is a block diagram showing an example of experiment system of application of the invention, and Fig. 3 is a diagram showing results of experiments of the relation of average bit error rate and reception CNR in the experiment systems in the drawings.

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Attorney: Suguru Kusano, patent attorney

Fig. 1

1 Voice signal input terminal

2 Facsimile signal input terminal

3 Data signal input terminal

4 Signal processing circuit

4a Error correction coding circuit

4d Time diversity circuit

5 Switch

6 Transmitter

7 Transmission antenna

9 Receiver

10 Switch

11 Signal processing circuit

12 Voice signal output terminal

13 Facsimile signal output terminal

14 Data signal output terminal

Fig. 2

1 Voice signal

2 Facsimile signal

3 Data signal

7 Transmission antenna

21 Transmitter-receiver

22 Reception diversity antenna

Fig. 3

Average bit error rate

Reception CNR (central value)

Voice

Facsimile

Data

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⑫公開特許公報（A）　昭63-28145

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⑤発明の名称　無線通信方式

㉑特願　昭61-172487
㉒出願　昭61(1986) 7月21日

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明細書

1.発明の名称

無線通信方式

2.特許請求の範囲

(1) 要求される伝送品質を異にする複数のサービスを無線通信により提供する方式であって、

上記各サービスに対し同一送信機により同一送信電力で無線通信を行い、

上記サービスの要求される伝送品質に応じてその要求が厳しい程、大きい改善効果が得られる伝送特性改善処理をそのサービス信号に対して施すことを特徴とする無線通信方式。

3.発明の詳細な説明

「産業上の利用分野」

この発明は複数サービスを提供する無線通信方式、特に移動通信方式に適する無線通信方式に関する。

「従来の技術」

移動通信において複数サービス（例えば音声、ファクシミリやデータ通信等）を提供しようとする場合、それらに要求される伝送品質（たとえばビット誤り率）が異なることが想定される。

移動通信では通常サービス領域の中心に固定通信網と接続されている無線基地局を設置し、そのサービス領域内を自由に移動する移動局はその無線基地局を介して固定通信網と接続される。移動局が通信できる範囲（ゾーン半径と呼ぶ）は、通信に要求される伝送品質と基地局／移動局の送信電力によって決まる。

一般には、ファクシミリやデータ通信では音声通信より厳しい伝送品質が要求されるため、音声通信に対して送信電力を設定したシステムにおいて同一の送信機、受信機を用いてファクシミリやデータ通信のサービスを受けようとすると、サービス領域の中心付近を除いてファクシミリやデータ通信のサービスを利用者が受けることが出来ない。そのため、音声通信が可能な全領域でファクシミリやデータ通信を品質良く行うためには、ファクシミリやデータ通信時には送信電力を大きくしなければならないことになる。

サービスごとに送信電力を制御することは比較的面倒になり、また送信電力を大にすると同一周波数を使用する無線通信システムの距離を離すことになり、従って周波数利用率が悪くなる、特に移動無線では同一周波数を用いるサービス領域の距離を離す必要があり周波数の有効利用が悪くなる。

この発明の目的は伝送品質を異にする複数のサービスの提供を同一の地域において同一送信機により同一送信電力で可能とする無線通信方式を提供することにある。

「問題点を解決するための手段」

この発明によれば同一送信機により同一送信電力で、要求される伝送品質が異なるサービスの信号を伝送し、そのサービスの信号をその要求される伝送品質に応じて異なる伝送特性改善処理を施し、この場合要求される伝送品質が厳しい程、大きい改善効果が得られるようにする。

このようにして全てのサービスに対して例えば同一のゾーン半径及び送信電力のもとで通信ができる。

「実施例」

第1図はこの発明の実施例を説明するための移動通信システムの例を示す。音声信号入力端子1、ファクシミリ信号入力端子2、データ信号入力端子3はそれぞれ伝送特性改善のための信号処理回路4を介してスイッチ5と接続される。この実施例では伝送特性改善技術として誤り訂正符号化及び時間ダイバーシチを用いる場合であって、信号入力端子1，2，3はそれぞれ信号処理回路4内の誤り訂正符号化回路4a，4b，4cにそれぞれ接続され、誤り訂正符号化回路4a，4b，4cの出力側は時間ダイバーシチ回路4d，4e，4fにそれぞれ接続され、これら時間ダイバーシチ回路4d，4e，4fはスイッチ5を介して送信機6に切替え接続される。送信機6の送信信号は送信アンテナ7より電波として送信される。

その電波は受信アンテナ8にて受信されて受信機9へ供給される。受信機9の出力側はスイッチ10を介して伝送特性改善のための信号処理回路11内の音声信号、ファクシミリ信号、データ信号と対応した回路の何れかに切替え接続される。信号処理回路11には音声信号出力端子12、ファクシミリ信号出力端子13、データ信号出力端子14が接続されている。

音声信号入力端子1には符号化された音声信号が入力される。その符号化音声信号は誤り訂正符号化回路4aによりチェックビットが付加された後、時間ダイバーシチ回路4dにより同一信号が複数回時間を隔てて送出される（時間ダイバーシチの動作については特願昭56-191814を参照）。ファクシミリ信号、データ信号に関しても同様にそれぞれ誤り訂正符号化回路4b，4c、時間ダイバーシチ回路4e，4fを通り、スイッチ5に入力される。スイッチ5は音声信号、ファクシミリ信号、データ信号のうちいずれか一つを選択して送信機6へ供給し、その信号は送信機6で搬送波を変調して送信アンテナ7より送信される。

その送信信号はアンテナ8で受信され、受信機9でベースバンド信号に復調復号された後、信号処理回路11に入力される。信号処理回路11は信号処理回路4の各処理の逆を行う回路であって音声信号、ファクシミリ信号、データ信号ごとにそれぞれ設けられ、それぞれ復調復号信号に対し時間ダイバーシチ処理の後、誤り訂正符号化処理が行われ、音声信号は音声信号出力端子12に、ファクシミリ信号はファクシミリ信号出力端子13に、データ信号はデータ信号出力端子14より出力される。

この場合、この発明では音声信号、ファクシミリ信号、データ信号ごとに訂正能力の異なる訂正符号及びブランチ数の異なる時間ダイバーシチを行い、つまり要求される伝送品質が高い程、誤り訂正符号の訂正能力を高め、時間ダイバーシチのブランチ数を増加する。例えば音声信号よりもファクシミリ信号の方を誤り訂正符号の訂正能力を高めかつ時間ダイバーシチのブランチ数を増加する。

このようにして異なる伝送品質を要求する複数のサービスを同一の送信電力、同一のゾーン半径

のもとで提供することが出来る。

なお伝送品質の要求に応じて誤り訂正符号の訂正能力のみ又は時間ダイバーシチのブランチ数のみを異ならしてもよい。

「発明の効果」

次にこの発明の効果を具体例について示す。音声信号として3kHzのアナログ信号をAPC-AB(適応予測・適応ビット割当)符号化した信号を、ファクシミリ信号としてG3の4.8kb/sの信号を、データ信号として2.4kb/sの信号を考え、これらに対する要求伝送品質をそれぞれ10^{-2}, 10^{-4}, 10^{-5}と仮定する。フェージング対策として2ブランチ空間ダイバーシチ(2SD)を用いると、音声信号については基地局／移動局の送信電力が15W/3Wのとき、1.5GHz帯でゾーン半径が3km、サービス領域に対する周波数割当てを9種類の周波数の組を繰返して実現される、しかしファクシミリ信号の場合は、送信電力を同一とすればゾーン半径1.4kmでサービス領域に対する周波数割当てを36種類の周波数の組を繰返して実現されることになる。

そこで第2図に示すように、入力端子1よりの音声信号はAPC-AB符号化回路15で符号化されると共にビット選別誤り訂正符号化(BSFEC)され、その符号化音声信号は16kb/sでスイッチ5へ出力される。ファクシミリ信号は誤り訂正符号化回路4bで誤り訂正符号化した後、時間ダイバーシチ回路4eで2ブランチの時間ダイバーシチ(2TD)を行って16kb/sでスイッチ5へ供給した。つまり時間ダイバーシチは2ブランチであるから、その1ブランチでは8kb/sが出力され、その3(8−4.8)kb/sが誤り訂正ビットに用いられる。端子3のデータ信号は誤り訂正符号化回路4cで誤り訂正符号化した後、時間ダイバーシチ回路4fで4ブランチの時間ダイバーシチ(4TD)を行ってスイッチ5へ16kb/sで供給した。送受信機21でGMSK(BbT=0.25)変調して送信した。つまり無線区間での伝送速度を16kb/sとした。受信は2ブランチ空間ダイバーシチアンテナ22で受信し、送受信機21で周波数検波2ビット積分検出方式で復調し、スイッチ5を通じて符号化回路15、時間ダイバーシチ回路4e,4fの何れかへ供給して復号した。

40Hzのレイリーフェージングの存在下における2ブランチ空間ダイバーシチのみを用いた場合(2SD)、2ブランチ空間ダイバーシチと2ブランチ時間ダイバーシチと誤り訂正符号とを用いた場合(2SD−2TD−FEC)、2ブランチ空間ダイバーシチと4ブランチ時間ダイバーシチと誤り訂正符号とを用いた場合(2SD−4TD−FEC)のそれぞれの受信CNR(中央値)に対する平均ビット誤り率の実験測定結果を第3図に示す。

この第3図より受信CNRが10dB附近で、音声信号は2SDによって平均ビット誤り率10^{-2}が得られ、ファクシミリ信号は2SD−2TD−FECで平均ビット誤り率10^{-4}が得られ、データ信号は2SD−4TD−FECで平均ビット誤り率10^{-5}が得られる。つまり音声信号、ファクシミリ信号、データ信号について第2図に示すような伝送特性改善処理をそれぞれ行えば同一送信電力で、それぞれ要求された伝送品質が得られる。前記移動無線に適用すると、ゾーン半径が3km、サービス領域に対する周波数割当てを9種類の周波数の組を繰返すことで音声信号のみならず、ファクシミリ信号、データ信号の何れのサービスの提供も行うことができる。

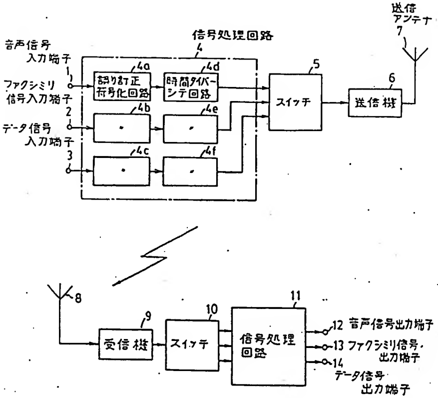
以上説明したように、この発明によれば同一送信機、同一送信電力で例えば音声通信が可能な地点でもファクシミリやデータ通信サービスが可能となり、利用者はサービスの違いを意識せずに複数サービスを受けることが出来る。この発明は移動通信のみならず一般の無線通信にも適用できる。

4.図面の簡単な説明

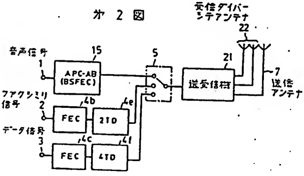
第1図はこの発明を適用した無線通信方式を示すブロック図、第2図はこの発明を適用した実験システムの例を示すブロック図、第3図は各図の実験システムについての平均ビット誤り率−受信CNRの関係の実験結果を示す図である。

特許出願人　日本電信電話株式会社
代理人　草野　卓

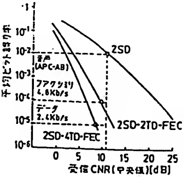
第1図


信号処理回路
4
音声信号
入力端子
1
ファクシミリ
信号入力端子
2
データ信号
入力端子
3
4a
誤り訂正
符号化回路
4d
時間ダイバーシテ回路
4b
4e
4c
4f
5
スイッチ
6
送信機
送信
アンテナ
7
8
9
受信機
10
スイッチ
11
信号処理
回路
12 音声信号出力端子
13 ファクシミリ信号
出力端子
14
データ信号
出力端子

第2図


受信ダイバーシテアンテナ
22
音声信号
1
15
APC-AB
(BSFEC)
5
21
送受信機
7
送信
アンテナ
ファクシミリ
信号
2
4b
FEC
4e
2TD
データ信号
3
4c
FEC
4f
4TD

第3図


平均ビット誤り率
1
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
10^{-6}
音声
(APC-AB)
ファクシミリ
4.8Kb/s
データ
2.4Kb/s
2SD
2SD-2TD-FEC
2SD-4TD-FEC
0
5
10
15
20
25
受信CNR(中央値)(dB)

P10568/P11166/P11667

No. 62-133842

SPECIFICATION

1. Title of the Invention

Multilevel quadrature amplitude modulation system

2. What is claimed is:

A multilevel quadrature amplitude modulation system, being used in a multilevel quadrature amplitude modulating apparatus comprising:

a differential-rotation object encoder (1) for encoding digital signals in plural systems being entered so as to remove the phase ambiguity of regenerative carrier,

a digital/analog converter (2) for converting the output of said differential-rotation object encoder into an analog signal, and

a modulator (3) for modulating the carrier in quadrature amplitude by the output of said digital/analog converter, comprising:

a signal point layout converter (4) for expanding the interval of signal points closest to the boundary of each quadrant,

wherein the bit error rate is improved.

3. Detailed Description of the Invention

[Summary]

In a multilevel quadrature amplitude modulation system,

a signal point layout converter is inserted between a differential-rotation object encoder for removing phase ambiguity of regenerative carrier and a digtal-to-analog converter, and the interval of signal points closest to the boundary of each quadrant is expanded, so that deterioration of bit error rate is improved.

[Industrial Field of Utilization]

The present invention relates to an improvement of multilevel quadrature amplitude modulation system used in digital microwave communications.

Recently, various digital microwave systems are realized, including the 64-level quadrature amplitude modulation system (hereinafter called 64QAM system), and to enhance the efficiency of use of frequency, there is a further multilevel trend, for example, 256QAM system.

The more advanced the multilevel trend, the severer is required the performance of the apparatus, and the apparatus is required, for example, to minimize the deterioration of bit error rate.

[Prior Art]

Fig. 3 is a block diagram of a conventional example, and Fig. 4 shows a signal point layout diagram of Fig. 3, both referring to the 256QAM system.

The numeral given beneath each point is data of second bit to fourth bit, and when the first bit numeral given in the upper right parentheses of each quadrant is added before the second bit, the corresponding data is obtained. For example, the data

at point A is 100 100, and when 11 at the upper right corner of the first quadrant is added, it actually shows 1100 1100.

Referring to Fig. 4, the operation in Fig. 3 is explained. First, in Fig. 3, the entered data of four bits and two systems (1ch, Qch) is given to the differential encoder 11 of the differential-rotation object encoder 1.

Herein, as disclosed in "Digital Microwave Communications" (Moriji Kuwabara, pp. 106-107, published by Project Center, March 1, 1985), in order to demodulate the correct data without knowing the absolute phase of transmission signal, the information is not placed on the position of signal point, but the information is placed on the transition of position.

That is, the summation operation of $y_i = x_i + y_{i-1}$ is converted by the differential encoder 11 into the original signal by differential operation of $x_i = y_i - y_{i-1} \ldots (x_i + y_{i-1}) - y_{i-1} = x_i$ in the reception side differential decoder (not shown), so that it is possible to demodulate without knowledge of absolute phase of the transmission signal.

Herein, y_i is the encoder output and x_i is the encoder input, and the summation operation and differential operation are paired operations, and both are combined and called differential conversion.

Consequently, the output of the differential encoder 11 is added further to the rotation object encoder 12, and, as shown in Fig. 12, the second-bit to fourth-bit codes are arranged so that equal codes of each quadrant may be at intervals of 90

degrees.

For example, in Fig. 4, signal point B of fourth quadrant and signal point B of first quadrant, and signal point C of first quadrant and signal point C of second quadrant are respectively at an interval of 90 degrees (except for the code of the first bit). Accordingly, when demodulating, if there is phase ambiguity of 90 × n degrees, such as 0, 90, 180 and 270 degrees n the phase of the reference carrier, no change occurs in the second to fourth bit, and the differential conversion may be done only on the first bit signal, and the signals of second to fourth bit are passed directly without being converted.

Herein, n is an integer.

When 1100 1100 is entered in the rotation object encoder 12, 1111 1111 is put out, and is converted into a maximum analog quantity in digital/analog converters 21, 22, and the carrier is modulated in quadrature amplitude in the modulator 3, and arranged at the position of point A.

Here, the input data are converted to analog amount corresponding to the respective positions of signal point layout diagram of Fig. 4, and disposed to positions as shown in Fig. 4, respectively.

[Problem that the Invention Is to Solve]

As shown in the signal point layout in Fig. 4, within a same quadrant, for example, if signal point D (000 010) of the lower three bits is mistaken to an adjacent bit 001 110, or 001 010, only one bit is wrong.

However, in the case of error over plural quadrants, a

multibit error occurs. For example, as shown in the column of "Number of errors when crossing quadrants" in Fig. 4, a maximum error of six bits may occur. This is a problem of deterioration of error rate.

[Means for Solving the Problem]

The problem is solved by the multilevel quadrature amplitude modulation system of the invention for improving the bit error rate, by disposing, as shown in Fig. 1, a signal point layout converter 4 in the multilevel quadrature amplitude modulating apparatus, and expanding the interval between the signals closest to the boundary of each quadrant.

[Operation of the Invention]

The invention has decreased the possibility of occurrence of error crossing over quadrants, by expanding the interval of signal points closest to the boundary of each quadrant.

That is, between a differential-rotation object encoder 1 and a digital/analog converter 2, a signal point layout converter 4 storing the signal point layout in Fig. 2, for example, a read-only memory is inserted, and the output of the rotation object encoder 12 is converted to the signal point layout in Fig. 2 and added to the digital/analog converter. As a result, the number of wrong signal points crossing quadrants is decreased, and the bit error rate is improved.

[Embodiment]

Fig. 1 is a block diagram of an embodiment of the invention, and Fig. 2 shows a signal point layout of Fig. 1, and the unit added in the embodiment of the invention is a signal point layout

converter 4.

Throughout the drawings, same reference numerals represent same components, and the 256QAM system is shown.

Referring now to Fig. 2 and Fig. 4, the operation in Fig. 1 is described below.

As shown in Fig. 1, the entered data of four bits and two systems (1ch, Qch) is converted into a 2-level signal as shown in signal point layout in Fig. 4 by the differential-rotation object encoder 1, and is added to the signal point layout converter 4. The signal point layout converter is composed of, for example, a read-only memory, which stores the data for converting the signal point layout in Fig. 4 into the signal point layout in Fig. 2, and the corresponding data is read out according to the output of the rotation object encoder as the address, and is added to the digital/analog converter 2 to be converted to an analog quantity, and the carrier is modulated in quadrature amplitude in the modulator 3, so that the 256QAM wave having the signal point layout as shown in Fig. 2 is obtained. As a result, the bit error rate is improved.

[Effects of the Invention]

As described in detail herein, the interval between signal points closest to the boundary of each quadrant is expanded, and hence the deterioration of bit error rate is improved.

4. Brief Description of the Drawings

Fig. 1 is a block diagram of an embodiment of the invention,

Fig. 2 is a signal point layout of Fig. 1,

Fig. 3 is a block diagram of a conventional example, and

Fig. 4 is a signal point layout of Fig. 3.

In the drawings,

1 is a differential-rotation object encoder,

2 is a digital/analog converter,

3 is a modulator, and

4 is a signal point layout converter.

Attorney: Sadakazu Igeta, patent attorney

Fig. 1 Block diagram of an embodiment of the invention.

1 Differential-rotation object encoder

3 Modulator

4 Signal point layout converter

11 Differential encoder

12 Rotation object encoder

Fig. 2 Signal point layout of Fig. 1.

Fig. 3 Block diagram of a conventional example.

3 Modulator

11 Differential encoder

12 Rotation object encoder

Fig. 4 Signal point layout of Fig. 3.

Number of errors when crossing quadrants

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⑤Int.Cl.⁴　　識別記号　　庁内整理番号　　⑬公開　昭和62年(1987)6月17日
H 04 L 27/00　　　　E－8226－5K

審査請求　未請求　発明の数 1　（全4頁）

⑭発明の名称　多値直交振幅変調方式

㉑特　願　昭60－274003
㉒出　願　昭60(1985)12月5日

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明細書

1．発明の名称

多値直交振幅変調方式

2．特許請求の範囲

再生キャリアの位相不確定が除去される様に、入力する複数系列のディジタル信号を符号化する差動・回転対象符号部(1)と、
該差動・回転対象符号部の出力をアナログ信号に変換するディジタル／アナログ変換部(2)と、
該ディジタル／アナログ変換部の出力で搬送波を直交振幅変調する変調器(3)とからなる多値直交振幅変調部において、
各象限の境界に最も接近している信号点の間の間隔を広げる信号点配置変換器(4)を設け、
ビット誤り率を改善する様にした事を特徴とする多値直交振幅変調方式。

3．発明の詳細な説明

（概要）

多値直交振幅変調方式において、再生搬送波の位相不確定が除去される差動・回転対象符号器とディジタル／アナログ変換器との間に信号点配置変換器を挿入して、各象限の境界に最も近い信号点の間の間隔を広げてビット誤り率の劣化を改善する様にしたものである。

（産業上の利用分野）

本発明は、ディジタルマイクロ波通信に使用する多値直交振幅変調方式の改良に関するものである。

近年、各種のディジタルマイクロ波方式、例えば64値直交振幅変調方式（以下64QAM方式と省略する）が実用化されているが、周波数利用効率を向上させる為に256QAM方式とより多値化の傾向にある。

しかし、多値化が進めば進む程、装置に対する要求性能が厳しくなるので、装置としては、例え

ばビット誤り率の劣化を出来るだけ少なくすることが必要である。

〔従来の技術〕

第3図は従来例のブロック図、第4図は第3図の信号点配置図を示すが、256QAM方式の場合を示す。

尚、各信号点の下に記載されている数字は第2ビット～第4ビットのデータで、各象限の右上の括弧内に記載されている第1ビットの数字を第2ビットの前に付加したものが対応するデータとなる。例えば、A点のデータは100 100 と記載されているが、第1象限の右上の1 1 を付加すると実際は1100 1100 を示す事になる。

さて、第4図を参照して第3図の動作を説明する。まず、第3図において、入力される4ビット2系列（Ich, Qch）のデータが差動・回転対象符号器1の中の差動符号器11に加えられる。

ここで、昭和60年3月1日企画センタ発行の桑原守二監修“デイジタルマイクロ波通信”p.106～107 で示される様に、送信信号の絶対位相を知らなくても正しいデータを復調できる様に、信号点の位置に情報を乗せず、位置の遷移に情報を乗せる。

即ち、差動符号器11でy_i・$x_i + y_{i-1}$の和分演算を、受信側の差動復号器（図示せず）でx_i・$y_i - y_{i-1}$=（$x_i + y_{i-1}$）$- y_{i-1} = x_i$の差分演算を行って原信号に変換する事により送信信号の絶対位相を知る事なしに復調できる。

尚、y_iは符号器出力，x_iは符号器入力を示し、和分演算と差分演算は対の操作であり、両者を合わせて差動変換と云う。

次に、差動符号器11の出力は更に回転対象符号器12に加えられ、第4図に示す様に、第2ビット～第4ビットの符号について、各象限の等しい符号が90度間隔になる様に配置される。

例えば、第4図の第4象限の信号点Bは第1象限の信号点Bと、第1象限の信号点Cは第2象限の信号点Cとそれぞれ90度の間隔になっている（第1ビットの符号を除く）。この為、復調の際に基準搬送波の位相に0，90，180，270度と90×n度の位相不確定があっても、第2～第4ビットに変化を生じないので、上記の差動変換は第1ビットの信号に対してのみ行えばよく、第2～第4ビットの信号は変換しないでそのまま通過させる。

ここで、nは整数を示す。

そこで、1100 1100 が回転対象符号器12に入力すると1111 1111 が出力され、デイジタル/アナログ変換器21,22 で最大のアナログ量に変換され、変調器3で搬送波を直交振幅変調してA点の位置に配置される。

以下、入力データは第4図の信号点配置図のそれぞれの位置に対応するアナログ量に変換され、第4図に示す様な位置にそれぞれ配置される。

〔発明が解決しようとする問題点〕

ここで、第4図の信号点配置図に示す様に、同一象限内で、例えば下位3ビットの信号点D(000 010）が誤りのビット001 110.又は 001 010に誤っても1ビットしか誤らない。

しかし、象限を越えて誤る時は多ビットの誤りを生ずる。例えば、第4図の“象限を横切る際の誤り数”の欄に示す様に最大6ビット誤ることがある。この為、誤り率が劣化すると云う問題点がある。

〔問題点を解決する為の手段〕

上記の問題点は、第1図に示す如く、多値直交振幅変調部に信号点配置変換器4を設け、各象限の境界に最も接近している信号点の間の間隔を広くしてビット誤り率を改善する様にした本発明の多値直交振幅変調方式により解決される。

〔作用〕

本発明は、各象限の境界に最も接近している信号点の間の間隔を広くする事により、象限を越えて誤りが発生する可能性を減少する様にした。

即ち、差動・回転対象符号器1とデイジタル／アナログ変換器2との間に第2図の信号点配置を記憶した信号点配置変換器4、例えばリード・オ

ンリ・メモリを挿入し、回転対象符号変換器12の出力を第2図の信号信号点配置になる様に変換して、デイジタル／アナログ変換器に加える様にした。そこで、象限を越えて誤る信号点の数が減るのでビット誤り率が改善される。

〔実施例〕

第1図は本発明の実施例のブロック図、第2図は第1図の信号点配置図を示し、本発明の実施例で付加された部分は信号配置変換器4である。

尚、全図を通じて同一記号は同一対象物を示し、256QAM方式の場合を示す。

そこで、第2図，第4図を参照しながら、第1図の動作を説明する。

第1図に示す様に、入力された4ビット2系列（Ich，Qch）のデータは差動・回転対象符号器1で第4図に示す様な信号点配置になる様な2値の信号に変換され信号点配置変換器4に加えられる。この信号点配置変換器は例えば、リード・オンリ・メモリで構成され、第4図の信号点配置を第2図の信号点配置に変換するデータが書込まれているので、回転対象符号器よりの出力をアドレスとして対応するデータが読出され、デイジタル／アナログ変換器2に加えられてアナログ量に変換された後、変調器3で搬送波を直交振幅変調して第2図に示す様な信号点配置を持つ256QAM波が得られる。これにより、ビット誤り率が改善される。

〔発明の効果〕

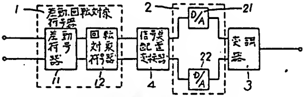
以上詳細に説明した様に、各象限の境界に最も接近している信号点の間の間隔を広くしたので、ビット誤り率の劣化が改善されると云う効果がある。

4．図面の簡単な説明

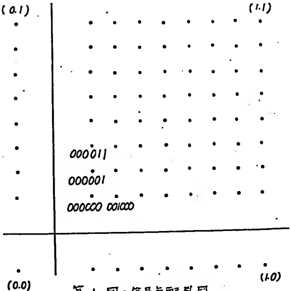
第1図は本発明の実施例のブロック図、
第2図は第1図の信号点配置図、
第3図は従来例のブロック図、
第4図は第3図の信号点配置図を示す。

図において、
1は差動・回転対象符号器、
2はデイジタル／アナログ変換器、
3は変調器、
4は信号点配置変換器を示す。

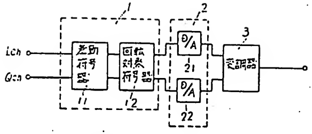
代理人　弁理士　井桁　貞一



本発明の実施例のブロック図
第1図



第1図の信号点配置図
第2図



従来例のブロック図
第 3 図

第3図の信号点配置図
第 4 図

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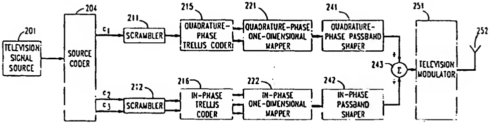
(54) Coding for digital transmission.

(57) Digital signals, such as digitized television signals, are subjected to a source coding step followed by a channel mapping step. The source coding step causes the television signal to be represented by two or more data streams while, in the channel mapping step, the mapping is such that the data elements of the various data streams have differing probabilities of being erroneously detected at the receiver. In proferred embodiments, a first one of the aforementioned data streams carries components of the overall television signal which are regarded as the most important--for example the audio, the framing information, and the vital portions of the video information, such as motion compensation information--and that data stream is mapped such that its data elements have the lowest probability of error. A second one of the data streams carries components of the overall television signal which are regarded as less important than those of the first data stream and that data stream is mapped such that its data elements have a probability of error that is not as low as those of the first data stream. In general, it is possible to represent the overall television signal with any number of data streams, each carrying components of varying importance and each having a respective probability of error. This approach allows a graceful degradation in reception quality at, for example, the television set location because as the bit error rate at the receiver begins to increase with increasing distance from the broadcast transmitter, the bits that represent proportionately less of the video information will be the first to be affected.

FIG. 2



EP 0 485 105 A2

Background of the Invention

The present invention relates to the transmission of digital data, including, but not limited to, the transmission of digital data which represents television signals.

It is generally acknowledged that some form of digital transmission will be required for the next generation of television (TV) technology, conventionally referred to as high definition television, or HDTV. This requirement is due mostly to the fact that much more powerful video compression schemes can be implemented with digital signal processing than with analog signal processing. However, there has been some concern about getting committed to an all-digital transmission system because of the potential sensitivity of digital transmission to small variations in signal-to-noise ratio, or SNR, at the various receiving locations.

This phenomenon--sometimes referred to as the "threshold effect"--can be illustrated by considering the case of two television receivers that are respectively located at 50 and 63 miles from a television broadcast station. Since the power of the broadcast signal varies roughly as the inverse square of the distance, it is easily verified that the difference in the amount of signal power received by the television receivers is about 2 dB. Assume, now, that a digital transmission scheme is used and that transmission to the receiver that is 50 miles distant exhibits a bit-error rate of 10^{-6}. If the 2 dB of additional signal loss for the other TV set translates into a 2 dB decrease of the SNR at the input of the receiver, then this receiver will operate with a bit-error rate of about 10^{-4}. With these kinds of bit-error rates, the TV set that is 50 miles away would have a very good reception, whereas reception for the other TV set would probably be very poor. This kind of quick degradation in performance over short distances is generally not considered acceptable by the broadcasting industry. (By comparison, the degradation in performance for presently used analog TV transmission schemes is much more graceful.)

There is thus required a digital transmission scheme adaptable for use in television applications which overcomes this problem. Solutions used in other digital transmission environments--such as the use of a) regenerative repeaters in cable-based transmission systems or b) fall-back data rates or conditioned telephone lines in voiceband data applications--are clearly inapplicable to the free-space broadcast environment of television.

Summary of the Invention

At the heart of our invention is the realization that a particular characteristic of prior art digital transmission systems is disadvantageous when carried over into, for example, the television transmission environment and that that characteristic lies at the crux of the problem. In particular, digital transmission systems have traditionally been engineered to provide about the same amount of protection against impairments to all the data elements--typically bits--that are transmitted over the communication channel. Such an approach is desirable when the digital transport mechanism is transparent to the user's data and no prior knowledge of the data's content is available--as is the case, for example, in voiceband data or digital radio applications. However, when all the bits are treated as equal, they are also all affected in the same way by changing channel conditions and the result may be catastrophic, as illustrated by the above example.

In accordance with the present invention, the shortcomings of standard digital transmission for over-the-air broadcasting of digital TV signals are overcome by a method comprising a particular type of source coding followed by a particular type of channel mapping--the latter being referred to herein as a catastrophe-resistant (C-R) mapping.

More specifically, the source coding step causes the television signal to be represented by two or more data streams while, in the channel mapping step, the mapping is such that the data elements of the various data streams havo differing probabilities of being erroneously detected at the receiver. In preferred embodiments, a first one of the aforementioned data streams carries components of the overall television signal which are regarded as the most important--as discussed in further detail hereinbelow--and that data stream is mapped such that its data elements have the lowest probability of error. A second one of the data streams carries components of the overall television signal which are regarded as less important than those of the first data stream and that data stream is mapped such that its data elements have a probability of error that is not as low as those of the first data stream. In general, it is possible to represent the overall television signal with any number of data streams, each carrying components of varying importance and each having a respective probability of error. This approach allows a graceful degradation in reception

quality at the TV set location because as the bit error rate at the receiver begins to increase with increasing distance from the broadcast transmitter, it will be the bits that represent the less important TV signal information that will be the first to be affected.

The invention is not limited to television signals but, rather, can be used in virtually any environment in which it is desired to provide different levels of error protection to different components of the intelligence being communicated.

Brief Description of the Drawing

In the drawing,

FIG. 1 is a block diagram of a transmitter embodying the principles of the invention, illustratively in the context of a four-dimensional channel mapping scheme for HDTV;

FIG. 2 is a block diagram of another transmitter embodying the principles of the invention in the context of a two-dimensional channel mapping scheme for HDTV, this scheme including trellis coding;

FIG. 3 is a block diagram of a receiver for transmitted signals transmitted by the transmitter of FIG. 1; and

FIGS. 4-11 are signal constellation maps useful in explaining the principles of the invention.

Detailed Description

Before proceeding with a specific description of the transmitters of FIGS. 1 and 2 and the receiver of FIG. 3, it will be helpful to first consider the theoretical underpinnings of the invention.

First off, it should be noted that the various digital signalling concepts described herein (with the exception, of course, of the inventive concept itself) are all well known in, for example, the digital radio and voiceband data transmission (modem) arts and thus need not be described in detail herein. These include such concepts as multidimensional signalling using N-dimensional signal constellations, where N is some integer; trellis coding; scrambling; passband shaping; equalization; Viterbi, or maximum-likelihood, decoding; etc. These concepts are described in such U.S. patents as U.S. 3,810,021, issued May 7, 1974 to I. Kalet et al.; U.S. 4,015,222, issued March 29, 1977 to J. Werner; U.S. 4,170,764, issued October 9, 1979 to J. Salz et al.; U.S. 4,247,940, issued January 27, 1981 to K. H. Mueller et al.; U.S. 4,304,962, issued December 8, 1981 to R. D. Fracassi et al.; U.S. 4,457,004, issued June 26, 1984 to A. Gersho et al.; U.S. 4,489,418, issued December 18, 1984 to J. E. Mazo; U.S. 4,520,490, issued May 28, 1985 to L. Wei; and U.S. 4,597,090, issued June 24, 1986 to G. D. Forney, Jr.--all of which are hereby incorporated by reference.

Turning now to the drawing, FIG. 4 depicts a standard two-dimensional data transmission constellation of the type conventionally used in digital radio and voiceband data transmission systems. In this standard scheme--conventionally referred to as quadrature-amplitude modulation (QAM)--data words each comprised of four bits are each mapped into one of 16 possible two-dimensional signal points. The constellation is thus labelled "Standard 16-QAM". Each signal point has an in-phase, or I, coordinate on the horizontal axis and has a quadrature-phase, or Q, coordinate on the vertical axis. Note that, on each axis, the signal point coordinates are ± 1 or ± 3 so that the distance between each point and each of the points that are horizontally or vertically adjacent to it is the same for all points--that distance being "2".

(The process of mapping the data words into particular signal points is referred to herein as "channel mapping" and the signal points are sometimes referred to as "channel symbols".)

Now consider the 16-point constellation of FIG. 5, which embodies the principles of the invention. The difference between this constellation and that of FIG. 4 is the relative distance between the different signal points.

Specifically, since the distance between all the adjacent points in FIG. 4 is the same, essentially the same probability of error is provided for all the bits which the signal points represent. (Transmission errors arise when, as the result of noise, phase jitter and various other channel phenomena/impairments, a transmitted signal point is displaced from its original position in the constellation to such an extent that it appears at the receiver that a different signal point was transmitted.) On the other hand, the distance between adjacent points in FIG. 5 is not the same for all the points. Specifically, the minimum distance between points within a particular quadrant in FIG. 5 is $d = \sqrt{2}$, and the minimum distance between points in adjacent quadrants is twice this amount, that is $2d = 2\sqrt{2}$. Thus, the probability of making an error in the receiver in identifying in which quadrant the transmitted point was located is smaller than the probability of making an error in identifying which point within that quadrant was the actual point. This results from the fact that the minimum distance between signal points representing different values of the data elements of the first data stream--e.g., the minimum distance ($2\sqrt{2}$) between the points in the first quadrant

representing the first-stream dibit 00 from those in the second quadrant representing the first-stream dibit 01--is greater than the minimum distance between the signal points representing the different values of the data elements of the second data stream--e.g., the minimum distance ($\sqrt{2}$) between the point in the first quadrant representing the second-stream dibit 00 and the point in that same quadrant representing the second-stream dibit 01.

Assume now that two out of the four bits of each transmitted data word need more protection from error than the other two bits because they are more important than the other two bits. This is achieved in accordance with the invention by using those two, more important bits to select one of the four quadrants (as indicated by the circled dibits in FIG. 5), and using the other two bits to select one of the four points within each quadrant, as indicated by the dibits next to each point. Since the probability of not correctly identifying the quadrant of the transmitted signal point is smaller than the probability of not correctly identifying the signal point itself, the desired protection is thereby achieved.

More generally stated, the constellation is divided into groups of signal points and each group is divided into subgroups, each of the latter being comprised of one or more signal points. At least one data element, e.g., bit, from each data word to be mapped identifies the group from which is to come the signal point that is to represent that data word, and at least one other data element identifies the subgroup within that group. If the subgroup contains more than one signal point, then further data elements are used to ultimately identify a particular one of those signal points (to which end the subgroup may be further divided into sub-subgroups). In accordance with the invention, a)the groups and subgroups are arranged such that the probability of the receiver erroneously determining which group a transmitted signal point is from is less than the probability of the receiver erroneously determining which subgroup it is from, and b) the data elements that identify the group represent information that is more important than the information represented by data elements that identify the subgroup.

A generic version of the constellation of FIG. 5 is shown in FIG. 6 in which the coordinate values, instead of being at $\pm\sqrt{2}$ and $\pm 2\sqrt{2}$, are $\pm\alpha$ and $\pm\beta$. It will also be appreciated that the constellations are not limited to any particular size, i.e., number, of signal points. For example, a standard 64-QAM constellation--represented by its upper right quadrant--is shown in FIG. 7, and a generic 64-point constellation embodying the principles of the invention and affording three different levels of protection is shown in FIG. 8.

Before proceeding, it is useful to make some formal definitions. As noted above, channel mapping in accordance with the invention is referred to herein as catastrophe-resistant (C-R) mapping. In general, a $(n_1, n_2, \cdots, n_k ; m)$ C-R mapping will be a mapping that provides the first (best) level of protection to n_1 bits; the second level of protection to n_2 bits; and so forth. The last entry in the mapping identification is a reminder of the total number m of information bits that are transmitted, that is: $m = n_1 + n_2 + \cdots + n_k$. With this definition, each of the C-R mappings shown in FIGS. 5 and 6 is a (2, 2 ; 4) mapping. FIG. 8 is an example of a 64-point (2, 2, 2 ; 6) mapping (represented by its upper right quadrant); and an example of a 16-point (1, 2, 1 ; 4) mapping is shown in FIG. 9. Finally, notice that standard QAM mappings of the type shown in FIGS. 4 and 7 can be considered as (m; m) C-R mappings.

We now briefly discuss the kind of trade-offs that are possible in the design of C-R mappings. First, we will assume that the power in the transmitted signal is subject to an average power constraint. Let a_i and b_i denote the I and Q discrete signal point levels, and assume that these signal points are uncorrelated. The average power constraint then requires that

$$\sum_i a_i^2 + \sum_i b_i^2 = constant \qquad (1)$$

for all the signal point level scenarios under consideration. Now let

$$SNR_{n_i}$$

denote the amount of SNR required to achieve a certain performance for the bits with the i^{th} level of protection. The change in the amount of SNR required by these bits to achieve that level of performance compared to a standard (m ; m) mapping is then defined by

$$\Delta SNR_{n_i} \equiv SNR_{n_i} - SNR_m , \qquad (2)$$

where SNR_m is the amount of SNR required by the (m ; m) mapping to achieve the same performance. (This is the mapping that provides the same amount of protection to all the bits.) With the expressions in (1) and (2), we get the following relationships for the (2, 2 ; 4) mapping shown in FIG. 6:

$$\beta = \sqrt{10-\alpha^2} \quad \Delta\ SNR_{n_1} = -20\log_{10}(\alpha) \quad \Delta\ SNR_{n_2} = -20\log_{10}\frac{\beta-\alpha}{2} , \qquad (3)$$

where the incremental SNRs are expressed in dB. Using α in FIG. 6 as a parameter in (3), we can first determine the value of β and then the incremental SNRs. Some computed values are given in Table I.

Table I - Trade-Offs for the (2, 2 ; 4) Mapping

α	β	$\Delta\ SNR_{n_1}$	$\Delta\ SNR_{n_2}$
1	3	0	0
1.1	2.965	- 0.83	0.61
1.2	2.926	- 1.58	1.28
1.3	2.883	- 2.28	2.03
$\sqrt{2}$	$2\sqrt{2}$	- 3	3
1.5	2.784	- 3.52	3.85
1.6	2.728	- 4.08	4.98

In order to give some meaning to the entries in Table I, we consider some specifics, such as the case $\alpha = \sqrt{2}$, which corresponds to the signal constellation shown in FIG. 5. The incremental SNRs for the two most protected bits are given in the third column. For the case under consideration, the incremental SNR for these bits is equal to -3 dB. Thus, for a given probability of error, these bits can tolerate an SNR that is 3 dB smaller than the SNR that would be required for a standard 16-point QAM system. On the other hand, as can be seen from the fourth column, the least protected bits would require three more dB of SNR in order to achieve the same performance as the standard QAM system.

The trade-off that has been achieved in the previous example may seem quite brutal: On one hand, we decrease the sensitivity to noise by 3 dB for the first two bits; and then, on the other hand, we increase this sensitivity by the same 3 dB for the other two bits. Such a clean trade-off rarely happens, as should be apparent from the other entries in Table I. For example, for $\alpha = 1.2$ more robustness against noise is gained by the most protected bits than is lost by the least protected bits. This is the kind of behavior to be sought in the design of efficient C-R mappings.

The invention is not limited to two-dimensional constellations but, indeed, can be implemented with N-dimensional constellations where $N \geq 2$. Indeed, an increase in the number of dimensions gives more flexibility in the design of efficient mappings. One way of implementing multidimensional C-R mappings with a QAM system is to use different two-dimensional C-R mappings in successive signal point intervals. As an example, a four-dimensional constellation can be created by concatenating all of the possible two-dimensional signal points from the (2, 2 ; 4) mapping of FIG. 5 with all of the possible two-dimensional signal points from the (1, 2, 1 ; 4) mapping of FIG. 9, as explained hereinbelow.

It is easily shown that such a mapping procedure provides a (3, 2, 3 ; 8) four-dimensional C-R mapping. Specifically, the greatest spacing between points in the constellation of FIG. 5 is the distance $2\sqrt{2}$, which is the smallest distance between the points in one quadrant and those in another. That same greatest spacing separates the upper and lower halves of the constellation of FIG. 9. Thus the highest level of protection can be achieved for three bits--two bits selecting a quadrant from the FIG. 5 constellation--as indicated by the circled dibits in FIG. 5--and a third bit selecting one of the two (upper and lower) halves of the constellation of FIG. 9--as indicated by the circled bits in FIG. 9. The next largest spacing is the distance between the columns in the constellation of FIG. 9, the smallest such distance being 2. Thus the second-highest level of protection is achieved for two bits, which select one of the four columns from the constellation of FIG. 9, as indicated by the squared-in dibits in FIG. 9. Finally, the smallest spacing is the distance $\sqrt{2}$ which, in the constellation of FIG. 5, is the smallest distance between the points within a quadrant and in the constellation of FIG. 9 is the smallest distance between the points within a column. Thus the lowest level of protection is again achieved for three bits--two bits selecting a point within the selected quadrant of the FIG. 5 constellation--as indicated by the dibits next to each point in FIG. 5--and a third bit selecting one of the two points contained within the selected half and selected column of the constellation of FIG. 9--as indicated by the single bit next to each point in FIG. 9.

It will thus be appreciated, by way of example, that the 8-bit word 01110100 would result in the selection of the four-dimensional signal point made up of the concatenation of point A from FIG. 5 and point A' from FIG. 9. Specifically, the first and second bits, 01, select the upper left quadrant of FIG. 5; the third bit, 1, selects the lower half of FIG. 9; the fourth and fifth bits, 10, select the second-from-right column of FIG. 9; the sixth and seventh bits, 10, select point A from the previously selected quadrant of FIG. 5; and the eighth bit, 0, selects point A' from the previously selected half and column of FIG. 9.

For this mapping, the SNR requirements for the two bits with the second level of protection are the same as the SNR requirements for the standard QAM signal constellation in FIG. 4. The three most protected bits and the three least protected bits have the SNR requirements that were derived in the previous section for the two-dimensional (2, 2 ; 4) mapping.

We are now in a position to consider the transmitter of FIG. 1.

Television signal source 101 generates an analog television signal which is passed on to source coder 104. The latter generates a digital signal in which at least one subset of the data elements represents information that is more important than the information represented by the rest of the data elements. Two examples of how such a signal might be generated are given hereinbelow.

The source-coded signal is illustratively C-R mapped, in accordance with the invention, using the four-dimensional mapping mentioned above in which each four-dimensional signal point is comprised of a two-dimensional signal point from the constellation of FIG. 5 concatenated with a two-dimensional signal point from the constellation of FIG. 9. In accordance with a feature of the invention, it has been recognized that it is desirable to preserve the distinctness of the bits which are to be accorded a particular level of protection by the mapping, notwithstanding any processing of the bits that may be necessary prior to their being mapped into four-dimensional signal points by four-dimensional mapper 121. Unless such distinctness is maintained, then, of course, it will not be possible to allocate the different levels of protection to the various data streams which represent the television signal.

In the present embodiment, in particular, it is desired to scramble the bits which comprise the digital signal in order to ensure a relatively uniform distribution of energy across the frequency band that the signal takes up. Accordingly, those bits are scrambled in three separate groups. Bits b_1, b_2 and b_3, which contain the most important information and are therefore to be accorded the highest level of protection, are scrambled by a first scrambler 111; bits b_4 and b_5, which contain the second-most-important information and are therefore to be accorded the second-highest level of protection, are scrambled by a second scrambler 112; and bits b_6, b_7 and b_8, which contain the least important amount of information and are therefore to be accorded the lowest level of protection, are scrambled by a third scrambler 113. (Scrambling is customarily carried out on a serial bit stream. Thus although not explicitly shown in FIG. 1, scramblers 111, 112 and 113 may be assumed to perform a parallel-to-serial conversion on their respective input bits prior to scrambling and a serial-to-parallel conversion subsequent thereto.)

The eight scrambled bits are applied in parallel in a four-dimensional mapper 121, mentioned above, which identifies a four-dimensional signal point to be generated using, for example, the bit-assignment scheme described above. Mapper 121 may be, for example, realized using table look-up. Conventional passband shaping and television modulation are then performed by passband shaper 141 and television modulator 151, respectively. The resultant analog television signal is then broadcast via antenna 152.

Turning now to the receiver of FIG. 3, the analog television signal is received by antenna 301, is subjected to conventional television front-end processing including, for example, demodulation in processing unit 311, and is converted to digital form by A/D converter 312. The signal is then equalized by passband channel equalizer 321 and passed on to detector 331. The latter stores information relating to the mapping--specifically, information indicative of the positions of the signal points of the constellation and the manner in which they are divided into groups and subgroups--and performs a so-called "slicing" operation on the equalized signal in order to form decisions as to what the transmitted signal points were in response to the stored information. Apart from having knowledge about the way in which the constellations of FIGS. 5 and 9 are configured pursuant to the invention, the detector is otherwise standard.

The 8-bit words output by detector 331 are descrambled by descramblers 341, 342 and 343, which respectively perform the inverse function of scramblers 111, 112 and 113 in the transmitter. A television signal formatted so as to be displayable by, for example, a CRT display is then generated from the descrambler outputs by picture signal generator 353. That signal is then applied to CRT display 360.

One more step of sophistication in the design of efficient C-R mappings can be achieved by adding redundancy to the signal constellations. Adding redundancy allows the usage of forward-error-correction coding, such as trellis coding. One of the issues with trellis coding is that its effect on the error rate of individual bits is not well understood. The published studies seem to have concentrated on the probability of error events, which is not easily related to the bit-error rate for trellis-coded systems. Nevertheless, even a simple example can show how more powerful C-R mappings can be obtained by using trellis coding.

Assume that we want to transmit three bits per signal point, and that one of the bits requires much more protection than the other bits. For a non-trellis-coded system, this can be done by using a (1, 2 ; 3) C-R mapping that has the signal constellation shown in FIG. 10. The most important bit defines the upper or lower half plane, and the other two bits define one of four possible points in each half plane. It is easily verified that the most valuable bit has 7 dB more margin against noise than the other two bits. We now assume that independent one-dimensional trellis codes are used along each axis in FIG. 10, thereby, in practical effect, increasing the distance between the rows and, independently, the distance between the points within a row. This leads to the signal constellation shown in FIG. 11. Specifically, one of the three bits is trellis-coded to become two bits which select one of the four rows in FIG. 11, and the other two bits are trellis-coded independently of the first bit to become three bits which select one of the eight columns in FIG. 11.

FIG. 2 shows a block diagram of a transmitter utilizing the constellation of FIG. 11. Television signal source 201 generates an analog television signal which is passed on to source coder 204. The latter generates a digital signal comprised of 3-bit binary data words c_1, c_2, c_3, in which it is assumed that bit c_1 is more important than the the other two bits c_2 and c_3. Bit c_1 is scrambled by a first scrambler 211, while bits c_2 and c_3 are scrambled by a second scrambler 212.

The output of scrambler 211 is trellis-encoded by quadrature-phase trellis coder 215, while the output of scrambler 212 is trellis-encoded by in-phase trellis coder 216. The 2-bit output of trellis coder 215 identifies one of the four rows of the FIG. 11 constellation, as described above. Those two bits are applied to quadrature-phase one-dimensional mapper 221, which generates an output identifying one of the four y-axis coordinates $\pm$ 1, $\pm$ 3. At the same time, the 3-bit output of trellis coder 216 identifies one of the eight columns of the FIG. 11 constellation. Those three bits are applied to in-phase one-dimensional mapper 222, which generates an output identifying one of the eight x-axis coordinates $\pm$ 0.5, $\pm$ 1.5, $\pm$ 2.5 and $\pm$ 3.5. Conventional passband shaping is then performed by quadrature-phase passband shaper 241 and in-phase passband shaper 242, whose outputs are combined in an adder 243. The resulting combined signal is then applied to television modulator 251, whose output analog signal is broadcast via antenna 252.

A specific receiver for the signal generated by the transmitter of FIG. 2 is not shown. Those skilled in the art will, however, be readily able to design such a receiver using standard building blocks similar to those used in FIG. 3, although in this case the detector stage preferably includes a maximum likelihood, or Viterbi, decoder in order to take advantage of the coding gain afforded by the trellis codes.

With this kind of trellis coding, we can decrease the sensitivity to noise by 3 dB for all the bits. For a probability of error of 10^{-6}, the most important bit then requires an SNR of about 11 dB, and the other two bits require an SNR of about 18 dB. (For simplicity, we assume here that the channel has a flat amplitude response.) If standard 8-point uncoded and 16-point trellis-coded signal constellations were utilized instead,

the SNR requirements would be 18 dB and 15 dB, respectively, for the same error rate. Designs for trellis-coded C-R mappings that are directly carried out in a multidimensional space should be even more powerful.

The foregoing merely illustrates the principles of the invention. For example, the invention is illustrated herein in the context of a digital TV transmission system. However, it is equally applicable to other types of digital transmission systems. Moreover, although particular constellations are shown herein, numerous other constellations, which may be of any desired dimensionality, can be used. For example, the various constituent two-dimensional C-R mappings that are used to provide higher-dimensional, e.g., four-dimensional, mappings may be used in unequal proportions. Alternatively, signal constellations with a different number of points may also be used in successive signal point intervals. All these possibilities provide a great flexibility for the design of efficient multidimensional C-R mappings. Additionally, it should be pointed out that four dimensions are naturally available in HDTV applications because of the possibility of using horizontal and vertical polarizations at the same time. Theoretically, this allows the simultaneous transmission of two independent QAM signals. Thus, for this application, there is an opportunity to implement multidimensional C-R mappings both in time (over different signal point periods) and in space (between polarizations).

Additionally, although a particular type of source coding is used in the illustrative embodiment hereof, it is envisioned that various other approaches to the digital representation of the television signal, i.e., other types of source coding, can be employed in order to give more protection to some of the transmitted bits. Such approaches might include the use of, for example, trellis/convolutional codes, BCH codes, Reed-Solomon codes, and/or concatenations of same. Disadvantageously, some channel mapping schemes may expand the bandwidth of the transmitted signal or may have potential synchronisation problems, and may not be cost effective. In any case, if these problems can be resolved, the present invention can always be combined with any such approachs since the latter operate on the bit stream. Also, it is envisioned that the source coding may well include other types of processing, such as any of various forms of television signal compression.

It may also be noted that although the invention is illustrated herein as being implemented with discrete functional building blocks, e.g., source coders, scramblers, etc., the functions of any one or more of those building blocks can be carried out using one or more appropriate programmed processors, digital signal processing (DSP) chips, etc.

It will thus be appreciated that those skilled in the art will be able to devise numerous and various alternative arrangements which, although not explicitly shown or described herein, embody the principles of the invention and are within its spirit and scope.

Claims

1. A method for communicating information
   CHARACTERIZED BY
   the steps of
   generating a digital signal representing the information, the digital signal being comprised of at least first and second data streams of data elements,
   channel mapping the digital signal, and
   transmitting the mapped signal over a communication channel,
   the mapping step being such that the probability of channel-induced error for the data elements of the first data stream is less than the probability of channel-induced error for the data elements of the second data stream.

2. The invention of claim 1
   CHARACTERIZED IN THAT
   said information is television signal information.

3. The invention of claim 1
   CHARACTERIZED IN THAT
   said mapping step includes the step of trellis coding the digital signal.

4. The invention of claim 1
CHARACTERIZED IN THAT
said generating step includes the steps of receiving the information, and source coding the information using a predetermined source code.

5. The invention of claim 1
CHARACTERIZED IN THAT
the mapping step comprises the step of
selecting a sequence of signal points from a predefined constellation of signal points to represent the data elements, the constellation being such that the minimum distance between signal points representing different values of the data elements of said first data stream is greater than the minimum distance between signal points representing the different values of the data elements of said second data stream.

6. The invention of claim 5
CHARACTERIZED IN THAT
said constellation is an N-dimensional constellation, where $N \geq 2$.

7. The invention of claim 5
CHARACTERIZED IN THAT
said generating step includes the steps of receiving the information, and source coding the information using a predetermined source code.

8. The invention of claim 7
CHARACTERIZED IN THAT
said generating step includes the further step of processing the source-coded information using at least a first predetermined processing algorithm, said processing being carried out for the data elements of said first data stream independently of the processing carried out for the data elements of said second data stream.

9. Apparatus for use in a digital transmission system of the type in which signal points from a predetermined signal point constellation representing respective associated data words are communicated from said transmitter over a communication channel to a receiver, said data words being comprised of individual data elements, said constellation being divided into groups of signal points and each of said groups being divided into subgroups of signal points, said apparatus including
means responsive to at least one of the data elements of each said data word for identifying which particular one of said groups includes the signal point associated with said data word and responsive to at least one other of the data elements of that word for identifying which particular one of the subgroups within said particular group includes the signal point associated with said data word, and
means for generating a signal representing a signal point from the identified subgroup and for applying that signal to said communication channel,
CHARACTERIZED IN THAT
said groups and subgroups are arranged such that the probability of said receiver erroneously determining which group a transmitted signal point is from is less than the probability of said receiver erroneously determining which subgroup it is from.

10. The invention of claim 9
FURTHER CHARACTERIZED IN THAT
the data elements that identify said particular group represent more important information than the data elements that identify said particular subgroup.

11. The invention of claim 9
CHARACTERIZED IN THAT
said information is television signal information.

12. The invention of claim 9
CHARACTERIZED IN THAT
said data words are trellis encoded data words.

13. The invention of claim 9
CHARACTERIZED IN THAT
said constellation is an N-dimensional constellation, where $N \geq 2$.

14. An arrangement for use in a receiver which receives intelligence communicated to said receiver by a transmitter, said transmitter including apparatus for a) generating a digital signal representing the intelligence, the digital signal being comprised of at least first and second data streams of data elements; b) channel mapping the digital signal using a predetermined signal constellation; and c) transmitting the mapped signal over a communication channel to the receiver, said mapping being such that the probability of channel-induced error for the data elements of the first data stream is less than the probability of channel-induced error for the data elements of the second data stream, said arrangement
CHARACTERIZED BY
means for receiving the transmitted signal, and
means for storing information relating to said signal constellation and for recovering said intelligence from the received signal in response to said stored information.

15. The invention of claim 14
CHARACTERIZED IN THAT
said intelligence is television signal information.

16. The invention of claim 15
CHARACTERIZED IN THAT
said mapping includes trellis coding of the digital signal and
FURTHER CHARACTERIZED IN THAT
said recovering means includes a maximum-likelihood decoder.

17. The invention of claim 14
CHARACTERIZED IN THAT
said mapping includes the selecting of a sequence of signal points from a predefined constellation of signal points to represent the data elements and
FURTHER CHARACTERIZED IN THAT
said stored information is indicative of the positions of the signal points of said constellation.

18. The invention of claim 17
CHARACTERIZED IN THAT
said constellation is an N-dimensional constellation, where $N \geq 2$.

19. The invention of claim 14 CHARACTERIZED IN THAT
said mapping includes the selecting of a sequence of signal points from a predefined constellation of signal points to represent the data elements, the constellation being such that the minimum distance between signal points representing different values of the data elements of said first data stream is greater than the minimum distance between signal points representing the different values of the data elements of said second data stream, and
FURTHER CHARACTERIZED IN THAT
the stored information is indicative of the positions of the signal points of said constellation.

20. A method for use in a receiver of a digital transmission system, said system being of a type in which signal points selected from a predetermined signal point constellation to represent respective associated data words are communicated from said transmitter over a communication channel to a receiver, said data words being comprised of individual data elements, said constellation being divided into groups of signal points and each of said groups being divided into subgroups of signal points, and said signal points being selected by following the steps of

a) identifying, in response to at least one of the data elements of each said data word, which particular one of said groups includes the signal point associated with said data word and, in response to at least one other of the data elements of that word, which particular one of the subgroups within said particular group includes the signal point associated with said data word, and

b) generating a signal representing a signal point from the identified subgroup and applying that signal to said communication channel, said groups and subgroups being arranged such that the probability of said receiver erroneously determining which group a transmitted signal point is from is less than the probability of said receiver erroneously determining which subgroup it is from,

said method
CHARACTERIZED BY
the steps of
receiving the transmitted signal points, and
recovering from the received signal points the data words represented thereby, said recovering being carried out in response to information stored, in said receiver, about said constellation and the manner in which it is divided into said groups and subgroups.

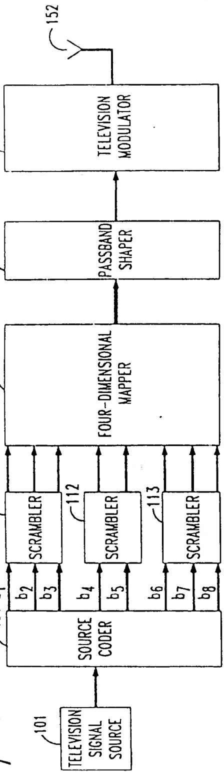
21. The invention of claim 20
FURTHER CHARACTERIZED IN THAT
the data elements that are used to identify said particular group represent more important intelligence than the data elements that identify said particular subgroup.

22. The invention of claim 20
CHARACTERIZED IN THAT
said information is television signal information.

23. The invention of claim 20
CHARACTERIZED IN THAT
said data words are trellis encoded data words.

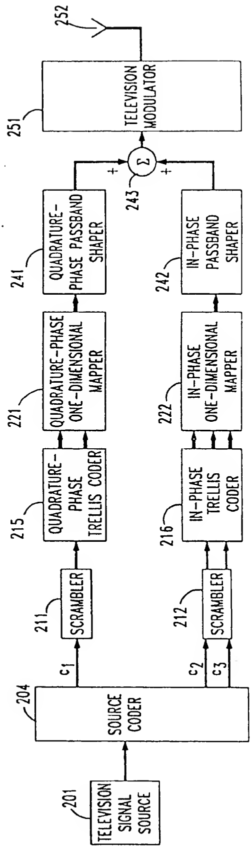
24. The invention of claim 20
CHARACTERIZED IN THAT
said constellation is an N-dimensional constellation, where $N \geq 2$.

FIG. 1



101
TELEVISION SIGNAL SOURCE
104
SOURCE CODER
b1
b2
b3
b4
b5
b6
b7
b8
111
SCRAMBLER
112
SCRAMBLER
113
SCRAMBLER
121
FOUR-DIMENSIONAL MAPPER
141
PASSBAND SHAPER
151
TELEVISION MODULATOR
152

FIG. 2



201
TELEVISION SIGNAL SOURCE
204
SOURCE CODER
c1
c2
c3
211
SCRAMBLER
212
SCRAMBLER
215
QUADRATURE-PHASE TRELLIS CODER
216
IN-PHASE TRELLIS CODER
221
QUADRATURE-PHASE ONE-DIMENSIONAL MAPPER
222
IN-PHASE ONE-DIMENSIONAL MAPPER
241
QUADRATURE-PHASE PASSBAND SHAPER
242
IN-PHASE PASSBAND SHAPER
243
Σ
+
+
251
TELEVISION MODULATOR
252

FIG. 3

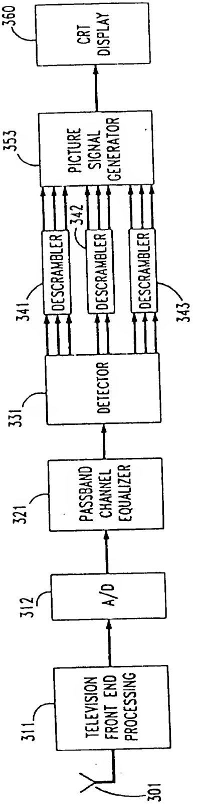
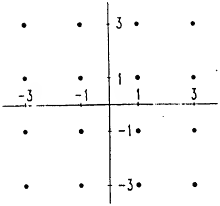
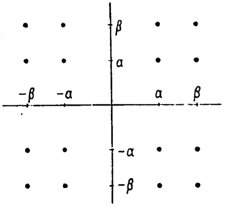

TELEVISION FRONT END PROCESSING
A/D
PASSBAND CHANNEL EQUALIZER
DETECTOR
DESCRAMBLER
DESCRAMBLER
DESCRAMBLER
PICTURE SIGNAL GENERATOR
CRT DISPLAY
301
311
312
321
331
341
342
343
353
360

FIG. 4



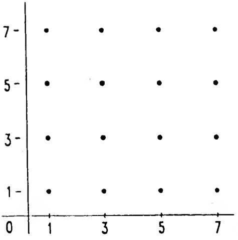
STANDARD 16-QAM

FIG. 6



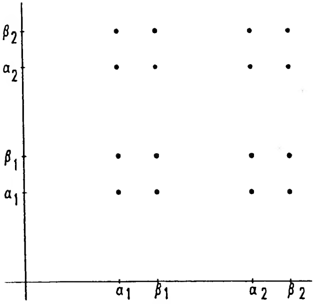
(2,2;4) C-R MAPPED 16-QAM

FIG. 7



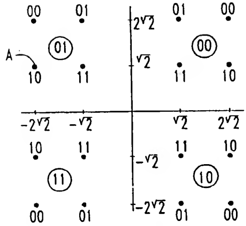
STANDARD 64-QAM

FIG. 8



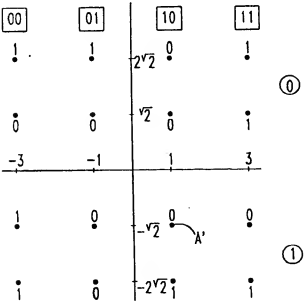
(2,2,2;6) C-R MAPPED 64-QAM

FIG. 5



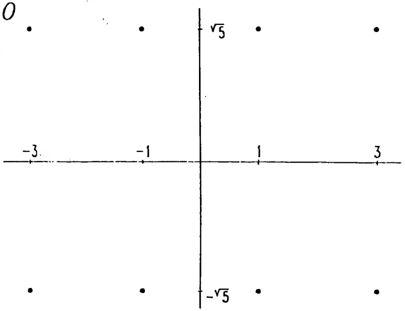
(2,2;4) C-R MAPPED 16-QAM

FIG. 9



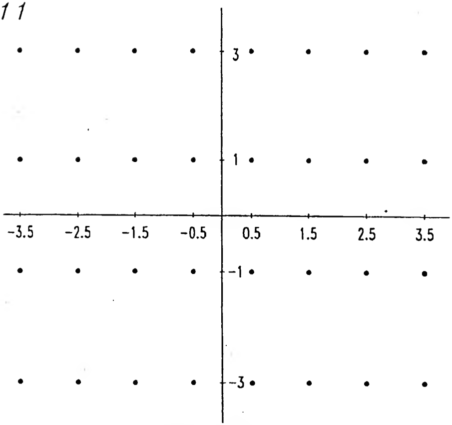
(1,2,1;4) C-R MAPPED 16-QAM

FIG. 10



(1,2;3) C-R MAPPED

FIG. 11



(1,2;3) TRELLIS-CODED C-R MAPPED

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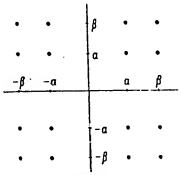
(54) Coding for digital transmission.

(57) Digital signals, such as digitized television signals, are subjected to a source coding step followed by a channel mapping step. The source coding step causes the television signal to be represented by two or more data streams while, in the channel mapping step, the mapping is such that the data elements of the various data streams have differing probabilities of being erroneously detected at the receiver. In preferred embodiments, a first one of the aforementioned data streams carries components of the overall television signal which are regarded as the most important--for example the audio, the framing information, and the vital portions of the video information, such as motion compensation information--and that data stream is mapped such that its data elements have the lowest probability of error. A second one of the data streams carries components of the overall television signal which are regarded as less important than those of the first data stream and that data stream is mapped such that its data elements have a probability of error that is not as low as those of the first data stream. In general, it is possible to represent the overall television signal with any number of data streams, each carrying components of varying importance and each having a respective probability of error. This approach allows a graceful degradation in reception quality at, for example, the television set location because as the bit error rate at the receiver begins to increase with increasing distance from the broadcast transmitter, the bits that represent proportionately less of the video information will be the first to be affected.

FIG. 6



(2,2;4) C-R MAPPED 16-QAM

EP 0 485 105 A3

European Patent Office

EUROPEAN SEARCH REPORT

Application Number

EP 91 31 0009
Page 1

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.5)
A	EP-A-0 282 298 (TANNER) * page 6, line 50 - page 7, line 28 * * page 7, line 37 - page 8, line 32; figures 5-7 * ---	1,9,14,20	H04L27/34 H04N7/13
O,A	EP-A-0 122 805 (CODEX CORPORATION) * page 4, line 27 - page 5, line 12 * * page 8, line 16 - line 28; figures 4,8,12,15 * * figures 17,20 * ---	1,9,14,20	
A	EP-A-0 302 622 (NORTHERN TELECOM LIMITED) * the whole document * ---	1,9,14,20	
A	US-A-4 346 473 (DAVIS) * the whole document * ---	1,9,14,20	TECHNICAL FIELDS SEARCHED (Int. Cl.5)
A	ELECTRONICS AND COMMUNICATIONS IN JAPAN. vol. 60-A, no. 12, 1977, NEW YORK US pages 18 - 25; S. HIRAKAWA ET AL.: 'Application of Multilevel Coding to Amplitude-and-Phase-Shift-Keying Systems' * page 19, left column, line 1, paragraph 2 - line 7, paragraph 4 * ---	1,9,14,20	H04L H04J H04N H03M
A	IEEE TRANSACTIONS ON INFORMATION THEORY. vol. 3, no. 4, October 1967, NEW YORK US pages 600 - 607; B. MASNICK ET AL.: 'On Linear Unequal Error Protection Codes' * the whole document * ---- -/--	1,9,14,20	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	31 JULY 1992	GRIES T.M.

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

European Patent Office

EUROPEAN SEARCH REPORT

Application Number

EP 91 31 0009
Page 2

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.5)
A	NTG FACHBERICHTE, Nummer 90, 'Bewegliche Funk-dienste', 1985, pages 178-185, VDE-Verlag, Frankfurt/Main, DE; L. KITTEL et al.: 'Verbundcodierung mit abgestuftem Fehlerschutz für digitale Funktelephonsysteme' * page 179, paragraph 3 - page 180, paragraph 4.1 *	1,9,14, 20	
A	ELECTRONICS LETTERS. vol. 20, no. 2, 1984, LONDON GB pages 62 - 63; E. L. CUSACK: 'Error Control Codes for QAM Signalling' * page 63, left column, paragraph 1 * * page 63, right column, paragraph 3 *	1,9,14, 20	

TECHNICAL FIELDS SEARCHED (Int. Cl.5)

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	31 JULY 1992	GRIES T.M.

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& : member of the same patent family, corresponding document

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OCC

(54) **Dispositif de transmission de données numériques à au moins deux niveaux de protection, et dispositif de réception correspondant.**

(57) Le domaine de l'invention est celui de la transmission de données numériques, notamment dans des canaux perturbés. Plus précisément, l'invention concerne la transmission dans un même canal de données nécessitant des niveaux de protection différents vis à vis des erreurs de transmission.

L'invention a notamment pour objectif d'optimiser l'utilisation du canal de transmission en affectant des techniques de transmission différenciées à des portions de données d'un même train numérique en fonction de niveaux de protection recherchés différents, contre les erreurs de transmission.

Cet objectif est atteint à l'aide d'un dispositif de transmission de données numériques à au moins deux niveaux de protection, du type assurant la répartition des données à transmettre sous forme d'éléments numériques dans l'espace temps-fréquence et l'émission de symboles constitués chacun d'un multiplex de N porteuses orthogonales modulées par un jeu desdits éléments numériques, et transmis simultanément, comprenant des moyens de codage canal comprenant au moins deux types de modulation (20_i, 21_i) et/ou au moins deux rendements de codage (30_i).

EP 0 448 492 A1

see EP-531046 A

RCPC, see COFDM
multi type modulation

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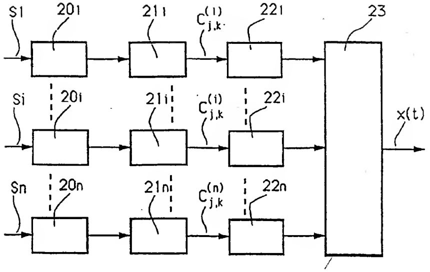

Fig. 2

DISPOSITIF DE TRANSMISSION DE DONNEES NUMERIQUES A AU MOINS DEUX NIVEAUX DE PROTECTION, ET DISPOSITIF DE RECEPTION CORRESPONDANT

Le domaine de l'invention est celui de la transmission de données numériques, notamment dans des canaux perturbés. Plus précisément, l'invention concerne la transmission dans un même canal de données nécessitant des niveaux de protection différents vis à vis des erreurs de transmission.

Les données transmises peuvent par exemple être des données sonores ou des données audiovisuelles (notamment télévision, visiophone, etc...), et plus généralement tout type de données numériques sur lesquelles il peut être intéressant, utile, ou en tout cas non néfaste d'effectuer une discrimination entre les éléments numériques sur un critère de niveau de protection minimale souhaité.

L'invention a pour arrière-plan technologique le système de radiodiffusion sonore numérique tel que décrit dans les demandes françaises 86 09622 du 2 juillet 1986, et 86 13271 du 23 septembre 1986 au nom des mêmes déposants. Le système de diffusion numérique présenté dans ces demandes de brevet antérieures est basé sur l'utilisation conjointe d'un dispositif de codage de canal, et d'un procédé de modulation par multiplexage de fréquences orthogonales (système COFDM : Coding Orthogonal Frequency Division Multiplex).

Le procédé de modulation proprement dit de ce système connu consiste à assurer la répartition d'éléments numériques constitutifs du signal de données dans l'espace fréquence-temps f-t et à émettre simultanément des jeux d'éléments numériques sur N voies de diffusion parallèles au moyen d'un multiplex de fréquences porteuses orthogonales. Ce type de modulation permet d'éviter que deux éléments successifs du train de données soit émis à la même fréquence. Ceci permet d'absorber la sélectivité fluctuante en fréquence du canal, en dispersant fréquentiellement, pendant la diffusion, les éléments numériques initialement adjacents.

Le processus de codage connu vise pour sa part à permettre de traiter les échantillons issus du démodulateur pour absorber l'effet de variation d'amplitude du signal reçu, due au processus de Rayleigh. Le codage est avantageusement un codage convolutif, éventuellement concaténé à un codage du type Reed-Solomon.

De façon connue, les éléments numériques codés sont de plus entrelacés, tant en temps qu'en fréquence, de façon à maximiser l'indépendance statistique des échantillons vis-à-vis du processus de Rayleigh et du caractère sélectif du canal.

Ce procédé est bien adapté à la diffusion de signaux numériques de haut débit (plusieurs mégabit/s) dans des canaux particulièrement hostiles comme sa première réalisation l'a montré dans le cadre de la radiodiffusion sonore numérique. Il permet notamment la réception de données numériques par des mobiles circulant en milieu urbain, c'est-à-dire en présence de parasites et de brouillage, et dans des conditions de multipropagation (processus de Rayleigh), engendrant un phénomène d'évanouissement (fading).

Toutefois, dans sa forme actuelle, ce procédé n'est pas utilisé de façon optimale : le même codage canal est utilisé pour l'ensemble des données à transmettre, avec la même protection contre les erreurs de transmission, quelle que soit l'importance des éléments de données.

Il est fréquent que des informations numériques destinées à être transmises dans le même canal aient des importances différentes. Ainsi, par exemple, dans le cas de signaux sonores, on sait que l'on peut tolérer un taux d'erreur d'environ 1% pour les bits les moins significatifs (LSBs), alors que les bits les plus significatifs (MSBs) exigent souvent un taux d'erreur inférieur à 10^{-6}. De même, dans un signal d'images, tous les coefficients transmis n'ont pas la même importance, en particulier d'un point de vue psychovisuel.

Il est clair que le taux d'erreur est lié notamment au type de codage utilisé, toutes conditions de réception étant égales par ailleurs, et en particulier aux procédés de corrections d'erreurs et aux redondances introduites. Il apparaît donc que le rendement du codage, en terme de débit, est lié au codage utilisé. En d'autres termes, plus le codage est fiable, plus son débit est faible.

Du point de vue du codage de canal seul, il est donc clair qu'un système de codage de canal protégeant uniformément le flot de données et basé sur la sensibilité aux erreurs de transmission des bits les plus significatifs est sous-optimal en terme d'efficacité spectrale (nombre de bits/s/Hz).

Il en résulte un codage de bonne qualité pour l'ensemble des bits, et donc un surcodage des bits peu importants, entraînant une perte de débit de transmission.

On connaît déjà des procédés réalisant une adaptation du codage de canal aux exigences du codage source. Il a notamment été proposé d'utiliser des codes convolutifs poinçonnés à rendement variable (RCPC) qui sont associés, à la réception, à un décodeur de Viterbi unique fonctionnant en décision douce. Ce procédé, décrit par R.V.Cox, N. Seshadri et C.-E. W. Sundberg dans "Combined subband source coding and convolutional channel coding" (Codage source par bandes et codage de canal convolutif combinés), ITG Tagung : Digitale Sprachverarbeitung, 26, 28-10-1988, Bad Nauheim, réalise la suppression, ou poinçonnage, périodique de certains bits du code source, lorsque le taux d'erreur maximum exigé le permet. Cependant ce type de codage reste lié à une modulation particulière, ce qui limite l'efficacité spectrale que l'on peut obtenir. Ainsi,

dans le cas d'un codage RCPC utilisé avec la modulation 4-PSK, il n'est possible d'atteindre au maximum qu'une efficacité spectrale strictement inférieure à 2. Par ailleurs, il ne semble pas possible d'utiliser cette technique dans de bonnes conditions avec des modulations à plus de quatre états de phase.

L'invention a pour objectif de pallier ces inconvénients.

Plus précisément, l'invention a pour objectif de fournir un dispositif de transmission numérique du type COFDM, optimisant le rendement de la transmission.

Un autre objectif de l'invention est de fournir un tel dispositif qui permette d'optimiser l'utilisation du canal de transmission en affectant des techniques de transmission différenciées à des portions de données d'un même train numérique en fonction de niveaux de protection recherchés différents, contre les erreurs de transmission.

Un objectif complémentaire de l'invention est de fournir un tel dispositif, exploitant la flexibilité et l'indépendance entre les porteuses du procédé COFDM.

Ces objectifs, ainsi que d'autres qui apparaîtront par la suite, sont atteints à l'aide d'un dispositif de transmission de données numériques à au moins deux niveaux de protection, du type assurant la répartition des données à transmettre sous forme d'éléments numériques dans l'espace temps-fréquence et l'émission de symboles constitués chacun d'un multiplex de N porteuses orthogonales modulées par un jeu desdits éléments numériques, et transmis simultanément, comprenant des moyens de codage canal comprenant au moins deux types de modulation et/ou au moins deux rendements de codage.

Ainsi, il est possible d'attribuer à chaque type de données à transmettre, en fonction du niveau de protection contre les erreurs requis, une modulation et/ou un rendement de codage adéquats.

Avantageusement, ledit multiplex de N porteuses est divisé en au moins deux jeux de porteuses, à chacun desdits jeux étant affecté un type de modulation différent et/ou un codage avec rendement de codage différent.

Dans ce cas, lesdits jeux de porteuses sont préférentiellement entrelacés selon l'axe fréquentiel, de façon à ce que chacun desdits jeux de porteuses bénéficie de l'indépendance en fréquence liée à la largeur de bande totale. En effet, on a intérêt à répartir les porteuses sur la plus grande largeur de bande possible, de façon à assurer une robustesse maximale vis à vis des perturbations sélectives en fréquence (notamment évanouissements).

Dans un autre mode de réalisation, le dispositif d'émission de l'invention comprend, pour au moins une desdites porteuses, des moyens de sélection entre au moins deux types de modulation et/ou entre au moins deux rendements de codage en fonction du débit de transmission et des perturbations affectant le canal.

Cela permet d'adapter de façon optimale le débit aux données à transmettre.

Avantageusement, dans ce second mode de réalisation, le dispositif d'émission comprend des moyens de génération de données d'assistance permettant, dans des récepteurs, de connaître pour chaque train de données numériques reçu, les types de modulation et/ou les rendements de codage sélectionnés correspondants.

Ces deux modes de réalisation peuvent également être mis en oeuvre simultanément, chaque jeu de porteuses pouvant utiliser au moins deux types de modulation et/ou deux rendements de codage, en fonction des données à transmettre.

De façon préférentielle, lesdits types de modulation sont des modulations de phase et/ou d'amplitude.

Dans un autre mode de réalisation avantageux, le dispositif de l'invention comprend, pour au moins une desdites porteuses, des moyens d'association optimale des éléments numériques codés aux états de la constellation de la modulation, selon la technique dite des modulations codées en treillis.

Pour permettre une démodulation cohérente, le dispositif comprend avantageusement des moyens d'insertion d'un motif de synchronisation fréquentiel récurrent dans le temps, permettant d'effectuer une démodulation cohérente dans lesdits récepteurs.

Préférentiellement, le dispositif d'émission de l'invention comprend au moins deux codeurs canal utilisant des polynômes générateurs identiques, de façon à permettre, dans les récepteurs, l'utilisation d'un même décodeur pour plusieurs trains de données codés par des codeurs distincts.

D'autres caractéristiques et avantages de l'invention apparaîtront à la lecture de la description suivante d'un mode de réalisation, donné à titre d'exemple illustratif et non limitatif, et des dessins annexés dans lesquels :

- la figure 1 présente des courbes du rapport de l'énergie par bit utile sur la densité spectrale de bruit en fonction de l'efficacité spectrale de différents modes de codage canal, dans les cas de canaux gaussien et de Rayleigh ;
- la figure 2 est un schéma de principe d'un dispositif d'émission selon l'invention ;
- la figure 3 est un synoptique d'une chaîne globale d'émission et réception selon l'invention, présentant les parties codage et décodage ;
- la figure 4 est un exemple d'entrelacement des jeux de porteuses, dans le cas de trois sources différentes,

du point de vue du niveau de protection contre les erreurs de transmission ;

– la figure 5 présente le synoptique détaillé d'un dispositif d'émission selon la figure 2, dans le cas d'une application à deux niveaux de protection ;

– la figure 6 présente le synoptique détaillé d'un dispositif de réception correspondant au dispositif d'émission de la figure 5.

Le dispositif de l'invention permet de résoudre de manière optimale le problème de transmission de différentes sources de données nécessitant des protections différentes. Il est basé sur l'utilisation du procédé COFDM. En effet, chacune des porteuses du multiplex OFDM est modulée de manière indépendante, ce qui permet de leur appliquer des modulations différentes.

Ainsi, à titre d'exemple, on peut envisager d'utiliser pour la transmission de données essentielles une modulation de phase à quatre états (4-PSK), et pour des données moins importantes, une modulation à 8 ou 16 états (8-PSK ou 16 PSK). Cette dernière modulation sera moins robuste que la première, mais chaque porteuse portera 1,5 (8-PSK) ou deux (16-PSK) fois plus d'informations, à technique de codage égale, ce qui entraînera une augmentation du débit final, sans modifier le taux d'erreur associé aux données essentielles.

Le débit global D d'informations binaires sortant d'un codeur source à transmettre sur un multiplex de N porteuses dans un canal de bande B donné, où B = N/ts, ts étant la durée d'un symbole élémentaire, peut s'écrire :

$$D = \sum_{i=1}^{n} D_i,$$

où n est le nombre de sources.

Si les différentes sources nécessitent des protections différentes vis à vis des erreurs de transmission, les valeurs de débit D_i peuvent être adaptées à chacune des sources.

Il est notamment possible, avec le procédé COFDM, de s'adapter à ce type de sources différenciées en agissant sur le rendement R_i du code associé à la source de débit D_i, par exemple en utilisant des techniques de poinçonnage.

La figure 1 présente deux courbes du rapport d'énergie par bit utile sur la densité spectrale de bruit (Eb/N_0), pour un taux d'erreur binaire de 10^{-4}, en fonction de l'efficacité spectrale (en bits/s/Hz) de la modulation, pour plusieurs types de modulation (4 PSK, 8-PSK, 16-QAM), dans des canaux gaussien et de Rayleigh. Pour un codage à 4 états de phase (4-PSK), on peut faire varier le rendement de 1/4 à 8/9, l'efficacité spectrale variant alors de 0,5 bit/s/Hz à près de 2 bits/s/Hz. Dans le même temps, le taux d'erreur augmente de façon importante, notamment dans le cas de canaux perturbés, du type canaux de Rayleigh sélectif. D'autre part, l'efficacité spectrale reste inférieure à 2 bits/s/Hz.

Il est donc plus intéressant, du point de vue efficacité en puissance, de passer à des constellations de modulation à plus grand nombre d'états associés à des procédés de codage adéquats selon le principe des modulations codées en treillis de Ungerboeck (MCT). On note par exemple qu'il faut mieux utiliser une modulation 8-PSK avec un rendement R = 2/3 (avec un codage MCT) qu'une modulation 4-PSK avec un rendement R = 8/9 (codage MCT).

Le système de l'invention permet également d'agir sur le type de modulation de chaque porteuse. Celle-ci sera caractérisée par le nombre de bits nbi porté par état de modulation. Une porteuse i comportera donc 2^{nbi} états.

Au débit D_i correspond donc en sortie du codeur un débit D_i/R_i, à répartir sur N_i porteuses modulées à 2^{nbi} états, avec les relations suivantes :

$$N = \sum_{i=1}^{n} N_i,$$

$$N_i = (D_i \,.\, ts) \,/\, R_i \,.\, nbi$$

On cherchera, pour être optimal, à adapter D_i et ts de façon à ce que N_i soit un nombre entier.

Si on applique le principe de la modulation codée en treillis décrit par C. Ungerboeck dans "Channel coding with multilevel phase signal" (codage de canal pour signal à plusieurs niveaux de phase), IEEE Transactions. Information theory, vol. I.T.28, janvier 1982, c'est-à-dire l'association optimale de mots codés de $n_i + 1$ bits sortant d'un codeur de rendement $R_i = n_i/(n_i + 1)$ aux états de la constellation de modulation 2^{n_i+1} états de manière à maximiser la distance entre signaux, on a également la relation suivante :

$$nbi = n_i+1$$

soit encore : $R_i \cdot nbi = n_i$

L'association optimale entre mots codés et états de modulation par le codage en treillis permet, à efficacité spectrale égale, un gain de codage important par rapport à un système de modulation ayant une constellation à 2^n états sans codage.

La figure 2 présente le schéma de principe d'un dispositif d'émission de n sources de données S1 à Sn selon l'invention, à n types de modulation, et donc n rendements de codage R_i différents.

Après l'opération 20_i (i variant de 1 à n) de codage de chaque série de données de débit D_i avec un rendement R_i, et d'allocation 21_i optimisée d'un état de modulation selon la méthode d'Ungerboeck, on obtient donc des symboles complexes $c^{(i)}j,k$, appartenant à un alphabet comportant 2^{ni+1} états. Les symboles $c^{(i)}_{j,k}$ sont ensuite entrelacés (22_i) en temps et en fréquence, puis, selon le procédé connu COFDM, subissent une transformée 23 de Fourier inverse, pour fournir le signal à émettre :

$$x(t) = \sum_{j=-\infty}^{+\infty} \sum_{i=1}^{n} \sum_{k \in I_i} C^{(i)}_{j,k} \ \varphi_{j,k}(t)$$

avec :

$\text{card}(I_i) = N_i$

$\varphi_{j,k}(t) = gk(t - jts)$ pour $0 \le t \le ts$

$gk(t) = e^{2i\pi f k t}$ pour $0 \le t \le ts$

0 ailleurs

$f_k = f_0 + k/ts$

i : indice de l'alphabet de modulation.

j : indice temporel des symboles

k : indice des porteuses.

A la réception, les échantillons complexes reçus après démodulation et transformée de Fourier discrète sont de la forme :

$$Y_{j,k}^{(i)} = H_{j,k} \cdot C_{j,k}^{(i)} + N_{j,k}$$

où $N_{j,k}$ représente un bruit gaussien complexe et $H_{j,k}$ la réponse du canal.

Chaque processus de décodage, associé à l'indice i, minimise alors, selon le critère de maximum de vraisemblance a posteriori, l'expression :

$$\Sigma\Sigma \mid \mid Y_{j,k}^{(i)} - H_{j,k}\, C_{j,k}^{(i)} \mid \mid / \, 2\, \sigma^2_{j,k}$$

où $\sigma^2_{j,k}$ est la variance de chaque composante de bruit gaussien complexe $N_{j,k}$.

L'invention ne se limite pas à l'utilisation de plusieurs types de modulation. Il est notamment possible d'utiliser également la technique de poinçonnage, ou toute autre technique permettant d'adapter le rendement de codage, avec un ou plusieurs types de modulation.

La figure 3 présente le synoptique général d'une chaîne d'émission et de réception selon l'invention, mettant en oeuvre plusieurs modulations, et la technique de poinçonnage RCPC.

Ce système réalise le codage différencié de cinq sources de données S1 à S5 nécessitant des niveaux de protection contre les erreurs de transmission distincts et décroissants.

Les trois premières sources de données S1, S2 et S3 sont codées selon une modulation 4-PSK 31_1, 31_2 et 31_3 avec des codes poinçonnés de rendements respectifs R1 = 1/4, R2 = 1/2 et R3 = 3/4 dans les codeurs 30_1, 30_2 et 30_3.

La source de données S4 est traitée par un codeur 30_4 en treillis à rendement R4 = 2/3, et une modulation 8-PSK 31_4, et la source de données S5 par un codeur 30_5 en treillis à rendement R5 = 5/6 et une modulation 64-QAM 31_5 (modulation d'amplitude en quadrature à 64 états), toutes deux traitées selon une technique de modulation en treillis.

Avantageusement, les polynômes générateurs des codeurs 30_4 et 30_5 sont identiques de façon que les données codées puissent être décodées, à la réception par un seul décodeur 37, si celui-ci est réalisé de manière suffisamment paramétrable.

Suivant la technique connue du codage COFDM, les différentes données codées sont soumises à une transformation 32 de Fourier rapide inverse (FFT^{-1}), puis émise dans le canal 33 de transmission.

A la réception, la démodulation 34 peut être soit différentielle (pour les modulations PSK), ainsi que cela est fait dans le système de radiodiffusion décrit dans les demandes de brevets français 86.09622 et 86.13271, déjà mentionnées, soit de façon cohérente, ainsi que cela est présenté dans la demande française 90.01492 du 06.02. 1990 au nom des mêmes déposants. Il est clair qu'une modulation QAM ne peut en revanche être démodulée que de façon cohérente.

Dans ce dernier cas, une méthode consiste à introduire dans le multiplex transmis un motif de synchronisation fréquentiel, récurrent dans le temps, permettant aux décodeurs de récupérer une référence de phase et/ou d'amplitude.

La partie réception comprend ensuite une transformation 35 de Fourier rapide (FFT), réalisant l'opération inverse de la FFT^{-1} 32, puis le décodage proprement dit.

Le choix de polynômes générateurs de codage identiques permet de limiter le nombre de décodeurs dans le récepteur.

Ainsi dans l'exemple donné, les trois sources S1, S2 et S3 pourront être décodées par le décodeur de Viterbi 36. Les deux sources S4 et S5 traitées par les deux décodeurs 30_4 et 30_5 en treillis ayant les mêmes polynômes, pourront aussi être décodées par le même décodeur 37 de Ungerboeck.

Le système COFDM utilise pleinement les deux dimensions temporelle et fréquentielle par son caractère large bande, et grâce à l'entrelacement temps-fréquence qui, associé au procédé de désentrelacement à la réception, permet d'obtenir à l'entrée du décodeur l'indépendance statistique maximale des échantillons successifs vis à vis des perturbations dues à la transmission.

Le procédé de l'invention permet de ne rien perdre en termes d'indépendance en fréquence, si l'on utilise un multiplexage fréquentiel optimal des différents peignes de porteuses associées aux différentes sources Di.

Pour cela, on entrelace les différents jeux de porteuses selon l'axe fréquentiel. Par exemple, dans le cas de trois sources différentes, le multiplexage pourra être tel que présenté en figure 4, pour les trois jeux de porteuses J1, J2, J3. Dans ce cas, chacun des trois jeux de porteuses bénéficie de l'indépendance en fréquence, liée à la largeur de bande totale.

Ainsi, le procédé de l'invention reste optimal pour chaque source Di en termes de puissance et d'efficacité spectrale.

L'approche décrite par Ungerboeck, définissant les bons codes et reposant sur l'association optimale des mots codés aux états de la constellation selon le critère de maximisation de distance entre signaux permet d'organiser les performances indépendamment pour chacune des sources Di.

On présente ci-dessous un exemple d'application chiffré, applicable notamment à la diffusion de séquences d'images réparties en deux trains d'éléments de données complémentaires b1 et b2, telle que décrit dans la demande de brevet conjointe de même date de dépôt au nom des mêmes déposants.

Dans ce cas, les paramètres de modulation et de codage sont fixés. Les dispositifs décrits sont néanmoins adaptables à un choix différent de ces paramètres.

On utilise un canal de transmission identique à celui utilisé dans le système de radiodiffusion sonore déjà réalisé. La largeur disponible du canal de transmission est B = N/ts = 7 MHz. La longueur des symboles Ts = 80 µs (comprenant la durée du signal utile ts = 64 µs et un intervalle de garde Δ = 16 µs). Le nombre de porteuses du multiplex N est alors égal à 448.

On se propose donc d'utiliser deux niveaux de protection différents vis à vis des erreurs de transmission :

- le premier niveau, associé au premier train de données b1, correspond au procédé qui a été utilisé lors de la première mise en oeuvre du codage COFDM dans le système de radiodiffusion connu. Ses paramètres sont les suivants :
  . modulation de phase à quatre états, démodulée en cohérent, soit une efficacité spectrale nb1 = 2 éléments binaires par Hertz (eb/Hz),
  . rendement de code R1 = 1/2,
  . nombre de porteuses du multiplex OFDM associé égal à N1.

Le débit utile transmis D1 est donc égal à :

$$D1 = nb_1 \times R_1 \times (N1/ts) \times (ts/Ts) = 2 \times (1/2) \times (N1/ts) \times (4/5).$$

Si on fixe N1 = 224, soit la moitié des porteuses disponibles on obtient un débit utile D1 = 2,8 Mbit/s.

- Le deuxième niveau de protection, associé au second train de données b2, fait appel aux techniques des modulations codées en treillis (Ungerboeck) en associant plus étroitement un code en treillis à une modulation à grand nombre d'états. Ses paramètres sont les suivants :
  . modulation de phase à huit états démodulée en cohérent, soit une efficacité spectrale nb_2 = 3 eb/Hz,
  . rendement du code R2 = 2/3,
  . le nombre de porteuses est N2.

Le débit utile transmis D2 bénéficiant de ce deuxième niveau de protection est égal à :

$$D2 = nb2 \times R2 \times (N2/ts) \times (ts/Ts) = 3 \times (1/2) \times (N2/ts) \times 4/5$$

Si on fixe N2 = 224, on obtient :

D2 = 5,6 Mbit/s.

Les deux trains de données b1 et b2 comprennent préférentiellement des données d'importances différentes, notamment selon un critère psychovisuel. Le procédé de l'invention permet de transmettre les données les plus pertinentes, correspondant au train b1, à l'aide d'un codage suffisamment robuste. Les données moins importantes du train b2 sont émises avec un niveau de protection contre les erreurs de transmission moins bon, ce qui est peu gênant, et avec un débit utile D2 double.

La figure 5 présente le schéma de principe d'un équipement d'émission correspondant à l'exemple décrit ci-dessus.

Les données 50 issues de la source sont séparées en deux trains binaires b1 et b2, de débits respectifs D1, D2, par un module de répartition 51.

Le premier train binaire b1 est traité de façon analogue à celle mise en oeuvre lors de la première réalisation du système COFDM. On effectue donc un codage convolutif 52, de rendement R1 = 1/2, puis un entrelacement temps/fréquence 53, suivi du codage binaire à signal 54. On obtient alors des données complexes $C_{j,k}$ qui sont traitées pour l'émission dans le module de modulation COFDM 56.

Le second train binaire b2 subit un codage 57 convolutif de type Ungerboeck, ou codage en treillis, à 2^{k-1} états, k étant la longueur de contrainte, et de rendement 2/3, puis une opération 58 d'association à chaque triplet de bits issus du codeur en treillis 57 un signal a_n de la constellation de la modulation de phase à états de phase, selon la méthode décrite par Ungerboeck sous le nom de "set partitioning" dans le document déjà mentionné.

Le signal a_n peut s'écrire : $a_n = e^{j(k\pi/8+\pi/16)}$, $k \in (0,\ldots,7)$.

Ce signal a_n est ensuite entrelacé en temps et fréquence (59) puis dirigé vers le module 56 de modulation COFDM.

De façon connue, ce module 56 réalise notamment une transformation de Fourier rapide inverse, sur des blocs de 512 mots complexes et une conversion numérique - analogique.

Les échantillons complexes résultants modulent ensuite une porteuse en phase et en quadrature, pour produire le signal 60 à émettre.

La figure 6 présente le schéma de principe de l'équipement de réception complet correspondant à l'émetteur décrit ci-dessus. Le signal reçu 60 est traité par le module 61 de démodulation COFDM, qui réalise notamment un filtrage de canal, une démodulation sur deux voies en quadrature par rapport à sa fréquence centrale, une numérisation et un traitement par un processeur de traitement du signal qui réalise une transformation de Fourier rapide (FFT).

Une fonction 62 d'estimation des porteuses du multiplex OFDM permet de réaliser la projection 63 sur les deux axes du plan complexe, à l'aide des mots de synchronisation fréquentielle, de façon à effectuer une démodulation cohérente.

Les deux trains b1 et b2 d'information sont ensuite décodés séparément. Le train b1 subit un désentrelacement temps-fréquence 64, puis est décodé par un décodeur de Viterbi 65. Le second train b2 est également désentrelacé (66) en temps et en fréquence, et décodé par un décodeur de Ungerboeck 67. Les données issues des deux décodeurs 65 et 67 sont ensuite regroupées, par un multiplexeur 68, de façon à fournir le signal 69 de données complet.

Dans l'exemple décrit, concernant la diffusion d'images numériques, il est possible de réaliser un second type de récepteur, plus simple, ne comprenant que le traitement lié au train d'information b1. Si la répartition entre les deux trains b1 et b2 est réalisée de façon judicieuse, il est en effet possible de reconstruire des images à l'aide du seul train b1. Evidemment, ces images seront de qualité inférieure, mais cependant, acceptable, notamment sur des écrans de petite taille.

Ces récepteurs utilisant le seul train b1, qui est codé de façon plus robuste, peuvent en particulier être utilisés dans des conditions de réception difficiles, telles que la réception dans des mobiles en milieu urbain.

Il est clair que l'exemple décrit ci-dessus n'est nullement limitatif de l'invention. Le nombre de sources d'information ou de trains de données à traiter avec des niveaux de protection distincts, peut être quelconque. Le niveau de protection peut être adapté soit en agissant sur le rendement de code utilisé, soit sur le type de modulation. Par ailleurs, l'invention ne s'applique pas seulement à la diffusion d'images numériques, mais également à la diffusion sonore, et, plus généralement, à la diffusion de tout type d'informations numériques. Elle permet de traiter de façon différenciée aussi bien des sous-ensembles d'un même programme, que des programmes complètement indépendants.

Dans un autre mode de réalisation, la modulation et/ou le rendement de codage affectés à chaque porteuse ou jeu de porteuses peut être variable, par exemple en fonction de l'importance des informations à transmettre

à chaque instant. De façon à permettre aux récepteurs de connaître la modulation et/ou le rendement sélectionnés, on génère à l'émission des données d'assistance. Ces données d'assistance doivent permettre au récepteur de fonctionner, en particulier dans le cas de réception sonore ou audiovisuelle, dès qu'il est mis en fonction. Cela peut être réalisé, par exemple, par l'affectation de certaines porteuses à la transmission des données d'assistance.

Revendications

1. Dispositif de transmission de données numériques à au moins deux niveaux de protection, du type assurant la répartition des données à transmettre sous forme d'éléments numériques dans l'espace temps-fréquence et l'émission de symboles constitués chacun d'un multiplex de N porteuses orthogonales modulées par un jeu desdits éléments numériques, et transmis simultanément, caractérisé en ce qu'il comprend des moyens de codage canal comprenant au moins deux types de modulation (20_i,21_i; 31_i) et/ou au moins deux rendements de codage (30_i).

2. Dispositif selon la revendication 1, caractérisé en ce que ledit multiplex de N porteuses est divisé en au moins deux jeux de porteuses (J1,J2,J3), à chacun desdits jeux étant affecté un type de modulation (20_i, 21_i ; 31_i) différent et/ou un codage (30_i) avec rendement de codage différent.

3. Dispositif selon la revendication 2, caractérisé en ce que lesdits jeux de porteuses (J1,J2,J3) sont entrelacés selon l'axe fréquentiel.

4. Dispositif selon l'une quelconque des revendications 1 à 3, caractérisé en ce qu'il comprend, pour au moins une desdites porteuses, des moyens de sélection entre au moins deux types de modulation (20_i,21_i ; 31_i) et/ou entre au moins deux rendements de codage (30_i) en fonction du débit de transmission et des perturbations affectant le canal.

5. Dispositif selon la revendication 4, caractérisé en ce qu'il comprend des moyens de génération de données d'assistance permettant, dans des récepteurs, de connaître pour chaque train de données numériques reçu, les types de modulation et/ou les rendements de codage (30_i) sélectionnés correspondants.

6. Dispositif selon l'une quelconque des revendications 1 à 5, caractérisé en ce que lesdits types de modulation (20_i,21_i ; 31_i) sont des modulations de phase et/ou d'amplitude.

7. Dispositif selon l'une quelconque des revendications 1 à 6, caractérisé en ce que lesdits niveaux de rendement de codage (30_i) sont obtenus par des moyens de poiçonnage à rendement variable du code source.

8. Dispositif selon l'une quelconque des revendications 1 à 7, caractérisé en ce qu'il comprend, pour au moins une desdites porteuses, des moyens (57) d'association optimale des éléments numériques codés aux états de la constellation de la modulation, selon la technique dite des modulations codées en treillis.

9. Dispositif selon l'une quelconque des revendications 1 à 8, caractérisé en ce qu'il comprend des moyens d'insertion d'un motif de synchronisation fréquentiel récurrent dans le temps, permettant d'effectuer une démodulation cohérente (62,63) dans lesdits récepteurs.

10. Dispositif selon l'une quelconque des revendications 1 à 9, caractérisé en ce qu'il comprend au moins deux codeurs canal (30_4,30_5) utilisant des polynômes générateurs identiques.

11. Dispositif de réception de données numériques transmises selon le dispositif de l'une quelconque des revendications 1 à 9, caractérisé en ce qu'il comprend autant de décodeurs (36,37) que ledit dispositif de transmission comprend de codeurs ayant des polynômes générateurs différents.

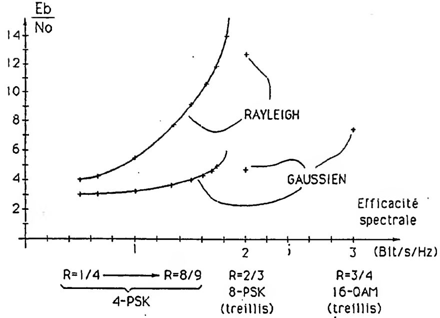
Eb/No
14
12
10
8
6
4
2
RAYLEIGH
GAUSSIEN
Efficacité spectrale
1
2
3
(Bit/s/Hz)
R=1/4 → R=8/9
4-PSK
R=2/3
8-PSK
(treillis)
R=3/4
16-QAM
(treillis)

Fig. 1

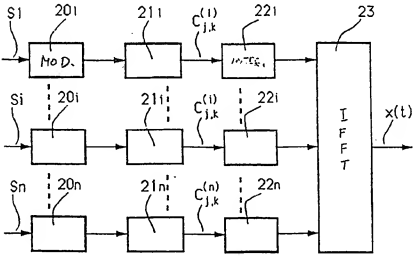
S1
201
211
221
23
MOD.
Si
20i
21i
22i
Sn
20n
21n
22n
IFFT
x(t)

Fig. 2

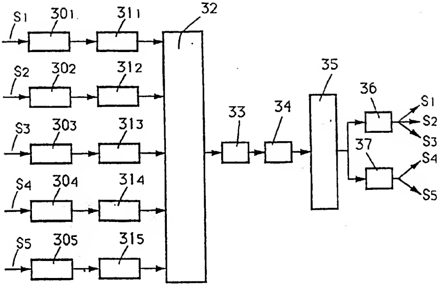
S1
301
311
32
S2
302
312
35
36
S3
303
313
33
34
S1
S2
S3
37
S4
S4
304
314
S5
305
315
S5

Fig. 3

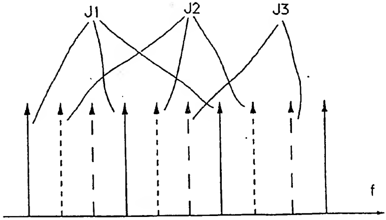
J1
J2
J3
f

Fig. 4

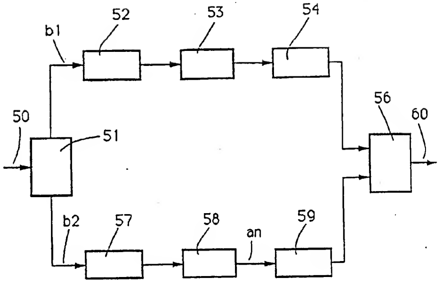
52
53
54
b1
56
50
60
51
b2
57
58
59
an

Fig. 5

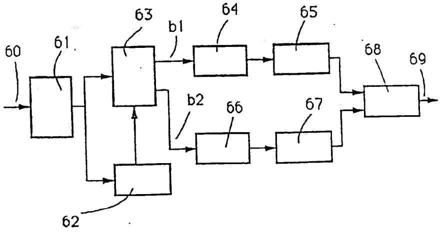
63
64
65
b1
60
61
68
69
b2
66
67
62

Fig. 6

Office européen des brevets

RAPPORT DE RECHERCHE EUROPEENNE

Numero de la demande

EP 91 46 0013

DOCUMENTS CONSIDERES COMME PERTINENTS

Catégorie	Citation du document avec indication, en cas de besoin, des parties pertinentes	Revendication concernée	CLASSEMENT DE LA DEMANDE (Int. Cl.5)
A,D	ITG TAGUNG: "Digitale Sprachverarbeitung", 26-28 octobre 1988, pages 23-30; R.V. COX et al.: "Combined subband source coding and convolutional channel coding" * Page 23, paragraphe 4 - page 24, paragraphe 2 * ---	1,2,7,8	H 04 L 5/06
A	EBU REVIEW - TECHNICAL, no. 224, août 1987, pages 168-190, Brussels, BE; R. LASSALLE et al.: "Principles of modulation and channel coding for digital broadcasting for mobile receivers" * Page 169, colonne de droite, alinéa 2 - page 170, colonne de droite, alinéa 4; page 178, colonne de droite, alinéas 5-7 * ---	1,3,6	
A	IEEE INTERNATIONAL CONFERENCE ON COMMUNICATIONS 1985, 23-26 juin 1985, CONFERENCE RECORD, vol. 2, pages 661-665, IEEE, New York, US; B. HIROSAKI et al.: "A 19.2 Kbps voiceband data modem based on orthogonally multiplexed QAM techniques" * Abrégé; page 663, colonne de gauche, paragraphe 1 * -----	8,9	DOMAINES TECHNIQUES RECHERCHES (Int. Cl.5) H 04 L

Le présent rapport a été établi pour toutes les revendications

Lieu de la recherche	Date d'achèvement de la recherche	Examinateur
LA HAYE	25-06-1991	SCRIVEN P.

CATEGORIE DES DOCUMENTS CITES

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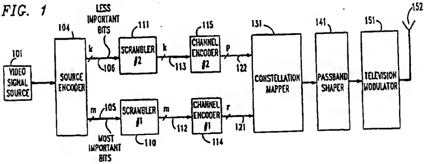
(54) Coded modulation with unequal levels of error protection.

(57) Digital signals, such as digitized television signals, are subjected to a source coding step in which a class of "most important" data elements represents a proportionately greater amount of the information to be communicated than the rest of the data elements. This is followed by a constellation mapping step which is carried out in such a way that those data elements have a lower probability of being erroneously detected at the receiver than the others. The constellation mapping step uses coded modulation in order to provide enhanced noise immunity for the "most important" data element class.

FIG. 1



EP 0 485 108 A2

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Background of the Invention

The present invention relates to the transmission of digital data, particularly the transmission of digital data which represents video signals.

It is generally acknowledged that some form of digital transmission will be required for the next generation of television technology, conventionally referred to as high definition television, or HDTV. This requirement is due mostly to the fact that much more powerful video compression schemes can be implemented with digital signal processing than with analog signal processing. However, there has been some concern about getting committed to an all-digital transmission system because of the potential sensitivity of digital transmission to small variations in signal-to-noise ratio, or SNR, at the various receiving locations.

This phenomenon--sometimes referred to as the "threshold effect"--can be illustrated by considering the case of two television receivers that are respectively located at 50 and 63 miles from a television broadcast station. Since the power of the broadcast signal varies roughly as the inverse square of the distance, it is easily verified that the difference in the amount of signal power received by the television receivers is about 2 dB. Assume, now, that a digital transmission scheme is used and that transmission to the receiver that is 50 miles distant exhibits a bit-error rate of 10^{-6}. If the 2 dB of additional signal loss for the other TV set translates into a 2 dB decrease of the SNR at the input of the receiver, then this receiver will operate with a bit-error rate of about 10^{-4}. With these kinds of bit-error rates, the TV set that is 50 miles away would have a very good reception, whereas reception for the other TV set would probably be very poor. This kind of quick degradation in performance over short distances is generally not considered acceptable by the broadcasting industry. (By comparison, the degradation in performance for presently used analog TV transmission schemes is much more graceful.)

There is thus required a digital transmission scheme adaptable for use in television applications which overcomes this problem. Solutions used in other digital transmission environments--such as the use of a) regenerative repeaters in cable-based transmission systems or b) fall-back data rates or conditioned telephone lines in voiceband data applications--are clearly inapplicable to the free-space broadcast environment of television.

An advantageous technique for overcoming the shortcomings of standard digital transmission for over-the-air broadcasting of digital TV signals--developed by employees of the assignee hereof--comprises a particular type of source coding step followed by a particular type of channel mapping step. More specifically, the source coding step causes the television signal to be represented by two or more data streams while, in the channel mapping step, the mapping is such that the data elements of the various data streams have differing probabilities of being erroneously detected at the receiver. Illustratively, a first one of the aforementioned data streams carries components of the overall television signal which are regarded as the most important--for example, the audio, the framing information, and the vital portions of the video information--and that data stream is mapped such that its data elements have the lowest probability of error. A second one of the data streams carries components of the overall television signal which are regarded as less important than those of the first data stream and that data stream is mapped such that its data elements have a probability of error that is not as low as those of the first data stream. In general, it is possible to represent the overall television signal with any number of data streams, each carrying components of varying importance and each having a respective probability of error. This approach allows a graceful degradation in reception quality at the TV set location because, as the bit error rate at the receiver begins to increase with increasing distance from the broadcast transmitter, it will be the bits that represent proportionately less of the TV signal information that will be the first to be so affected.

Summary of the Invention.

In accordance with the present invention, I have devised a scheme which implements the above-described overall concept of providing different levels of error protection for different classes of data elements generated by the source encoding step--but which provides enhanced noise immunity via the use of coded modulation, such as trellis-coded modulation.

In preferred embodiments of the invention, in particular, the symbols in a predetermined 2N-dimensional channel symbol constellation, $N \geq 1$, are divided into groups, each of which is referred to herein as a "supersymbol." During each of a succession of symbol intervals, a predetermined number of the most important data elements are channel encoded, and the resulting channel coded data elements identify a particular one of the supersymbols. The remaining data elements, which may also be channel encoded, are used to select for transmission a particular symbol from the identified supersymbol.

The approach as thus far described is similar in a general way to conventional coded modulation schemes in that the latter also divide the channel symbols into groups, typically referred to as

"subsets." However, the prior art subsets are formed under the constraint that the minimum Euclidean distance (hereinafter referred to as the "minimum distance") between the symbols in a subset is greater than the minimum distance between the symbols in the constellation as a whole. In accordance with the present invention, however, the minimum distance between the symbols of a supersymbol is the same as the minimum distance between the symbols in the constellation as a whole. It is this distance property which allows for greater amount of noise immunity for the most important data elements than for the other data elements, that immunity being optimized by keeping the minimum distance between supersymbols as large as possible--usually greater than the minimum distance of the constellation. Specifically, once the supersymbols are defined, it is possible to design codes for the most important data elements based on the distances between the supersymbols, i.e., as though each supersymbol were a conventional symbol in a conventional constellation. This being so, a particular degree of noise immunity can be achieved for the most important data elements that is greater than what can be achieved for the other data elements.

Indeed, a tradeoff is involved in that those other data elements suffer a coding loss, i.e., a somewhat lessened noise immunity. Importantly, however, the coding gain that can be achieved for the most important data elements is greater than that which can be achieved using conventional coded modulation schemes.

Brief Description of the Drawing

In the drawing,

FIG. 1 is a block diagram of a transmitter embodying the principles of the invention;

FIG. 2 is a block diagram of a receiver for transmitted signals transmitted by the transmitter of FIG. 1;

FIG. 3 depicts a prior art signal constellation;

FIG. 4 depicts a signal constellation illustratively used by the transmitter of FIG. 1;

FIG. 5 shows a bit assignment scheme for the constellation of FIG. 4;

FIG. 6 shows a type of trellis encoder that can be used in the transmitter of FIG. 1;

FIG. 7 is a table comparing the performance of the various illustrative embodiments of the invention disclosed herein;

FIG. 8 depicts an alternative signal constellation that can be used in the transmitter of FIG. 1;

FIGS. 9-11, when taken together, show another type of trellis encoder that can be used in the transmitter of FIG. 1;

FIG. 12 depicts yet another signal constellation that can be used in the transmitter of FIG. 1;

FIG. 13 shows another type of trellis encoder than can be used in the transmitter of FIG. 1;

FIG. 14 depicts yet another signal constellation that can be used in the transmitter of FIG. 1; and

FIG. 15 shows how a bit interleaver can be added to one of the channel encoders in the transmitter of FIG. 1 to provide enhanced impulse noise immunity.

Detailed Description

Before proceeding with a description of the illustrative embodiments, it should be noted that the various digital signaling concepts described herein--with the exception, of course, of the inventive concept itself--are all well known in, for example, the digital radio and voiceband data transmission (modem) arts and thus need not be described in detail herein. These include such concepts as multidimensional signaling using 2N-dimensional channel symbol constellations, where N is some integer; trellis coding; scrambling; passband shaping; equalization; Viterbi, or maximum-likelihood, decoding; etc. These concepts are described in such U.S. patents as U.S. 3,810,021, issued May 7, 1974 to I. Kalet et al.; U.S. 4,015,222, issued March 29, 1977 to J. Werner; U.S. 4,170,764, issued October 9, 1979 to J. Salz et al.; U.S. 4,247,940, issued January 27, 1981 to K. H. Mueller et al.; U.S. 4,304,962, issued December 8, 1981 to R. D. Fracassi et al.; U.S. 4,457,004, issued June 28, 1984 to A. Gersho et al.; U.S. 4,489,418, issued December 18, 1984 to J. E. Mazo; U.S. 4,520,490, issued May 28, 1985 to L. Wei; and U.S. 4,597,090, issued June 24, 1986 to G. D. Forney, Jr.--all of which are hereby incorporated by reference.

Turning now to FIG. 1, video signal source 101 generates an analog video signal representing picture information or "intelligence" which signal is passed on to source encoder 104. The latter generates a digital signal in which at least one subset of the data elements represents a portion of the information, or intelligence, that is more important than the portion of the information, or intelligence, represented by the rest of the data elements. Illustratively, each data element is a data bit, with m+k information bits being generated for each of a succession of symbol intervals. The symbol intervals are comprised of N signalling intervals, where 2N is the number of dimensions of the constellation (as described below). The signaling intervals have duration of T seconds and, accordingly, the symbol intervals each have a duration of NT seconds. In

embodiments using two-dimensional constellations, i.e., N = 1, then of course the signaling intervals and the symbol intervals are the same.

Of the aforementioned m+k information bits, the bits within the stream of m bits per symbol interval, appearing on lead 105, are more important than the bits within the stream of k bits per symbol interval, appearing on lead 106. Two examples of how one might generate a television signal of this type are given hereinbelow.

The bits on leads 105 and 106 are independently scrambled in scramblers 110 and 111, which respectively output m and k parallel bits on leads 112 and 113. (Scrambling is customarily carried out on a serial bit stream. Thus although not explicitly shown in FIG. 1, scramblers 110 and 111 may be assumed to perform a parallel-to-serial conversion on their respective input bits prior to scrambling and a serial-to-parallel conversion at their outputs.) In accordance with the invention, as described more fully hereinbelow, the respective groups of bits on leads 112 and 113 are extended to channel encoders--illustratively trellis encoders--114 and 115 which generate, for each symbol interval, respective expanded groups of the expanded r and p bits on leads 121 and 122, where r > m and p > k. The values of those bits jointly identify a particular channel symbol of a predetermined constellation of channel symbols (such as the constellation of FIG. 4 as described in detail hereinbelow). Complex plane coordinates of the identified channel symbol are output by constellation mapper 131, illustratively realized as a lookup table or as a straightforward combination of logic elements. Conventional passband shaping and television modulation are then performed by passband shaper 141 and television modulator 151, respectively. The resultant analog signal is then broadcast via antenna 152 over a communication channel, in this case a free-space channel.

In order to understand the theoretical underpinnings of the invention, it will be useful at this point to digress to a consideration of FIG. 3. The latter depicts a standard two-dimensional data transmission constellation of the type conventionally used in digital radio and voiceband data transmission systems. In this standard scheme--conventionally referred to as quadrature-amplitude modulation (QAM)--data words each comprised of four information bits are each mapped into one of 16 possible two-dimensional channel symbols. Each channel symbol has an in-phase, or I, coordinate on the horizontal axis and has a quadrature-phase, or Q, coordinate on the vertical axis. Note that, on each axis, the channel symbol coordinates are ± 1 or ± 3 so that the distance between each symbol and each of the symbols that are horizontally or vertically adjacent to it is the same for all symbols--that distance being "2". As a result of this uniform spacing, essentially the same amount of noise immunity is provided for all four information bits.

As is well known, it is possible to provide improved noise immunity without sacrificing bandwidth efficiency (information bits per signaling interval) using a coded modulation approach in which an "expanded" two-dimensional constellation having more than (in this example) 16 symbols is used in conjunction with a trellis or other channel code. For example, my above-cited '490 patent discloses the use of a 32-symbol, two-dimensional constellation together with an 8-state trellis code. That coded modulation scheme achieves approximately 4 dB of enhanced noise immunity over the uncoded case of FIG. 3, while still providing for the transmission of four information bits per signaling interval. Here, too, however, essentially the same amount of noise immunity is provided for all four information bits.

In accordance with the invention, the known noise immunity and bandwidth efficiency advantages of coded modulation are achieved while providing different levels of error protection to different classes of bits. Specifically, I have discovered that it is possible to achieve a level of error protection for a class of "most important" bits which is substantially greater than what can be achieved with the aforementioned conventional coded modulation approach. Indeed, the transmitter of FIG. 1 embodies the inventive concept, as will now be described in further detail.

The constellation used in the transmitter of FIG. 1 is illustratively the two-dimensional 64-symbol constellation shown in FIG. 4 (each symbol is represented as a dot in the figure). In accordance with the invention, the symbols in the signal constellation are divided into groups which I refer to as "supersymbols." Specifically, the constellation of FIG. 4 is divided into $2^r = 2^3 = 8$ supersymbols. Four of the supersymbols, labeled 000, 011, 100 and 111, are each comprised of eight contiguous channel symbols assigned to that supersymbol. The other four supersymbols, labeled 001, 010, 101 and 110, are each comprised of two non-contiguous groups, each comprised of four contiguous channel symbols. (The use of such two-group supersymbols allows the overall constellation to have, for example, better signal-to-noise ratio, lower peak-to-average power ratio and better symmetry than would otherwise be possible.)

In this example, m = k = 2. That is, 50% of the bits are in the class of most important bits. Each of encoders 114 and 115 adds one redundant bit, so that r = p = 3. The r = 3 bits on lead 121 identify one of the eight supersymbols and the p = 3 symbols on lead 122 select a particular one of the eight channel symbols within the identified

supersymbol. In accordance with an important aspect of the invention, the minimum distance between the symbols of a supersymbol–that distance being denoted d_2–is the same as the minimum distance between the symbols in the constellation as a whole. Indeed, it can be verified by observation that this criterion is satisfied in FIG. 4. Given this characteristic, increased noise immunity for the most important bits can be provided via appropriate selection of a) the codes implemented by encoders 114 and 115 and b) the ratio d_1/d_2–where d_1 is the minimum distance between the supersymbols. (The parameter d_1 is given by the minimum of the distances between all the pairs of supersymbols. In turn, the distance between any pair of supersymbols is the minimum distance between any symbol of one of the pair of supersymbols and any symbol of the other.)

Specifically, a coded modulation scheme can now be constructed for the most important bits as though the eight supersymbols were eight conventional symbols in a conventional constellation. (It is true that in a conventional constellation a symbol cannot be divided into halves, as is the case for supersymbols 001, 010, 101 and 110. However, for the purpose of coding design, one may treat each of the halved supersymbols as being located in only one of its two positions.) To design such a coded modulation scheme, the eight supersymbols are partitioned, as is conventional, into a predetermined number of subsets and an appropriate code is used to encode some of the most important input bits to generate a stream of coded output bits which define a sequence of subsets. The remaining most important input bits are then used to select a supersymbol from each identified subset. In this particular example, each subset contains only a single supersymbol, i.e., there are eight subsets, and all most important input bits--i.e, the two bits on lead 112–are encoded. Thus the identification of a particular subset also identifies a particular supersymbol. It is from this particular supersymbol that the symbol that is ultimately to be transmitted will be selected as a function of the other, or less important, bits.

Importantly, it will be appreciated that this approach achieves--by virtue of the partitioning and code selected–a particular degree of noise immunity for the most important data elements that is greater than what can be achieved for the less important data elements and, indeed, is greater than what can be achieved with conventional coded modulation, all other things being equal.

As noted above, the less important bits, on lead 113, are then used to select a particular symbol from the identified supersymbol for transmission. In preferred embodiments, this selection also involves the use of coded modulation wherein at least some of the less important bits are encoded to identify a particular subset of symbols within a supersymbol and, if the subset contains more than one symbol, any remaining bits are used to select a particular one of those symbols. (The arrangement of the symbols within a supersymbol should, of course, be chosen jointly with encoder 115 to maximize its coding gain.) Again in this example, there are eight subsets of symbols within each supersymbol, i.e., one symbol per subset, and both of the less important bits on lead 113 are encoded. Thus the three coded bits on lead 122 identify, at one and the same time, both a subset and a specific symbol from the earlier identified supersymbol.

A particular illustrative embodiment for both encoders 114 and 115 is shown in FIG. 6. (In this FIG., the boxes labelled "T" are T-second delay elements, the circles labelled "+" are exclusive-or gates, and the two-input gates are AND gates, one of which has one of its inputs inverted.) As noted above, the 3-bit output of encoder 114 identifies a particular supersymbol. Specifically, the bit values "110" output by encoder 114 on its three output leads (reading from top to bottom in FIG. 6) identifies the supersymbol labeled 110 in FIG. 4, and so forth for each of the seven other possible bit patterns. Additionally, the 3-bit output of encoder 115 selects a particular symbol within the identified supersymbol. In particular, the assignment of bit values to particular channel symbols within the supersymbols is shown in FIG. 5 for the upper right quadrant of the FIG. 4 constellation. The bit assignment scheme for the other three quadrants are arrived at by simply rotating FIG. 5. Thus, for example, the bit values "001" output by encoder 115 on its three output leads (reading from top to bottom) identifies the channel symbol labeled 001 in the identified supersymbol–there being one such symbol in each supersymbol.

Given the use of the particular trellis code implemented by the encoder of FIG. 6, various operational parameter tradeoffs can be achieved by varying the values of d_1 and d_2. Two possibilities for the constellation of FIG. 4 are shown in the table of FIG. 7. In particular, with d_1/d_2 = 2.5, a coding gain of 5.7 dB (measured, relative to an uncoded 16-QAM scheme such as shown in FIG. 3–which has the same bandwidth efficiency as the current example–at a block error rate of 10^{-3} for a block size of 1,000 bits) is achieved for the most important bits at a cost of a coding gain of -2.8 dB (i.e., a coding loss) for the less important bits. Alternatively, with d_1/d_2 = 3.5, a coding gain of 6.8 dB is achieved for the more important bits at a cost of a coding gain of -4.6 dB for the less important bits. The peak-to-average power ratio is about "2"

(as it is for all the examples described herein), which is comparable to that achieved with conventional uncoded modulation.

Turning now to the receiver of FIG. 2, the analog broadcast signal is received by antenna 201, is subjected to conventional television front-end processing including, for example, demodulation in processing unit 211, and is converted to digital form by A/D converter 212. The signal is then equalized by passband channel equalizer 221 and passed on parallel rails 222 and 223 to channel decoders 231 and 232. Each of the channel decoders is, illustratively, a maximum likelihood decoder, such as a Viterbi decoder. Specifically, the function of channel decoder 231 is to identify the most likely sequence of supersymbols, while the function of channel decoder 232 is to identify the most likely sequence of symbols, given that sequence of supersymbols. Thus, decoder 231 has stored within it information about the code used by channel encoder 114, while decoder 232 has stored within it information about the code used by channel encoder 115. Additionally, between the two of them these two decoders have stored within them information about the constellation being used and the manner in which the symbols are assigned to their respective supersymbols.

In channel decoder 231, the first step of decoding is to find the supersymbol or half supersymbol in each subset that is closest to the received symbol–such as the point denoted "A" in FIGS. 4 and 5. In this case, it will be remembered, there is only one supersymbol per subset. Channel decoder 231 assumes a specific single location in the signal space for each supersymbol or half supersymbol. Three such locations, denoted with a dashed "x", are shown in FIG. 4. The other locations are placed similarly. The distance between that supersymbol or half supersymbol and the received symbol is then determined. (The distance between the received symbol and a supersymbol or half supersymbol is the distance between the former and the previously defined location of the latter.) After this, decoding proceeds to find the most likely sequence of transmitted supersymbols in just the same way that a Viterbi decoder operates in a conventional coded modulation system to find the most likely sequence of conventional symbols.

The operation of channel decoder 232 will be explained with reference to FIG. 5. The first step is to rotate the received symbol by an integral multiple of 90 degrees so that the resulting symbol is always in, say, the so-called first quadrant, which is the quadrant depicted in FIG. 5. It is then determined whether the rotated symbol is closer to supersymbol 000, or one of the first-quadrant halves of supersymbols 001 and 101. After this, for each subset of symbols of the supersymbol or two supersymbol halves (in this decoding procedure, these two supersymbol halves are treated as if they belonged to the same supersymbol), the symbol that is closest to the rotated symbol is identified and the distances between them are calculated. This information is then used by channel decoder 232 to identify–for the purpose of recovering the less important bits–the most likely sequence of transmitted symbols. Alternatively stated, this determination of the most likely sequence of transmitted symbols is used only for purposes of extracting the less important bits. The more important bits are recovered from channel decoder 231 as described above.

An alternative way of realizing decoder 232 is to wait for decoder 231 to form its decision as to the identity of each supersymbol and then use this information in the recovery of the less important bits. (No rotation would be required in this case.) Such an approach has the potential advantage of allowing one to use a more complex code for the less important bits--and thereby achieve greater noise immunity for them–but at a cost of increased receiver processing delay.

Decoding in the case where multi-dimensional symbols are used–such as the four-dimensional examples described below–is carried out in a similar manner, as will be appreciated by those skilled in the art.

The bits output by decoders 231 and 232 are descrambled by descramblers 241 and 242, which respectively perform the inverse function of scramblers 110 and 111 in the transmitter. A video signal formatted so as to be displayable by, for example, a CRT display is then generated from the descrambler outputs by source decoder 253, thereby recovering the original video information, or intelligence. That signal is then applied to CRT display 260.

Numerous variations of the invention are possible. Consider, for example, the two-dimensional constellation of FIG. 8, which is comprised of four supersymbols each being comprised, in turn, of eight symbols. This constellation could be used in a system having m = 1, k = 2–i.e., the bandwidth efficiency is three information bits per signaling interval and the most important bits constitute 33.3% of the total--and in which each channel encoder introduces one redundant bit, i.e., r = 2) and p = 3. However, in order to increase the bandwidth efficiency, this same constellation can be used as the basis of a four-dimensional code which supports four information bits per signaling interval.

Specifically, the four-dimensional constellation is constructed by concatenating the constellation of FIG. 8 with itself so that each four-dimensional

symbol is comprised of a first point selected from the two-dimensional constellation concatenated with a second such point. (We herein use the word "point" to refer to an element of the two-dimensional constellation of FIG. 8, thereby differentiating it from the overall coded entity, which we consistently refer to herein as a "symbol," no matter what its dimensionality. We will use the term "superpoint" in a similar way.) For this four-dimensional case, m = 3, k = 5 information bits are input to channel encoders 114 and 115, respectively, for each symbol interval of duration 2T. This provides an average of four information bits per signaling interval (or eight information bits per symbol interval). The more important bits in this case constitute 3/(3+5) = 3/8 = 37.5% of the information bits.

FIG. 9 shows the structure of channel encoders 114 and 115 for this four-dimensional embodiment. Encoder 114 adds a single redundant bit to its 3-bit input to provide a pair of 2-bit outputs which respectively identify first and second superpoints from FIG. 8. The first point of the four-dimensional symbol to be transmitted is to be selected from the first such superpoint and the second point of the four-dimensional symbol to be transmitted is to be selected from the second such superpoint.

The less important bits are used to provide such selection. Specifically, encoder 115 adds a single redundant bit to the 5-bit input on lead 113 to provide two 3-bit outputs which, as just noted, respectively select specific points from the first and second superpoints identified by encoder 114.

Specific circuitry for carrying out the actual encoding within channel encoders 114 and 115 is shown in FIG. 10, the bit converter of which operates in accordance with the table of FIG. 11.

The relative performances achieved for this embodiment with various values of d_1/d_2 are shown in FIG. 7. Note that if one is willing to have the most important bits constitute a lower percentage of the total--37.5% in this embodiment compared to 50% for the first embodiment--a greater coding gain can be achieved for such bits.

A further characteristic of coded modulation schemes based on FIG. 8--which is independent of the dimensionality of the overall code--is that it allows for the use of coded modulation schemes which are expected to provide greater immunity against impulse noise for the most important bits than other constellations, such as that shown in FIG. 4 and FIG. 12 (the latter being described below). The reason is that the positions of the various superpoints relative to one another can be defined based solely on angular, as opposed to amplitude, information.

A further protection against impulse noise for the most important bits in coded modulation schemes based on constellations of the type of FIG. 8 can be achieved by rearranging the bits that are output by channel encoder 114 so that bits that are generated in proximity to one another by the encoder are separated from one another as much as possible, given that the system delay constraints are met. To this end, channel encoder 114 may include a bit interleaver, which performs such rearrangment, as shown in FIG. 15. (In the receiver, a complementary de-interleaver will, of course, be used--before channel decoder 231.) On the one hand, it can be shown that for coded modulation schemes in which the Euclidean distance between valid sequences of supersymbols is the same (with the possible exception of a scaling factor) as the Hamming distance betweeen the sequences of bits associated with those sequences of supersymbols--which is the case for the coded modulation scheme just described--such rearrangement of the bits does not degrade the performance of the code against additive white Gaussian noise. On the other hand, however, such rearrangement provides an enhanced immunity against impulse noise. This is a result of the bursty nature of impulse noise. (Enhanced impulse noise immunity can also be achieved for coded modulation schemes which do not meet the above criteria--such as the various other schemes disclosed herein--by rearranging the two-dimensional signal points before transmitting them. This approach is somewhat less effective, however, than when the bits are rearranged.)

As another alternative, consider, for example, the two-dimensional constellation of FIG. 12, which is comprised of eight supersymbols each being comprised, in turn, of four symbols. This constellation could be used in a two-dimensional signaling scheme having m = 2, k = 1--i.e., the bandwidth efficiency is three information bits per signaling interval and the more important bits constitute 66.7% of the total--and in which each channel encoder introduces one redundant bit, i.e., r = 3 and p = 2. As before, however, in order to increase the bandwidth efficiency, this same constellation can be used as the basis of a four-dimensional code. Here we would have m = 5 and k 3 for an average of four information bits per signaling interval. The more important bits in this case constitute 5/8 = 62.5% of the information bits. It will be appreciated that this embodiment is similar to that previously described except that the channel encoders for the most- and less-important bits are exchanged. Finally, it may be noted from FIG. 7 that the increased percentage of most important bits brings with it a decreased coding gain for those bits.

It is also important to note that the constellations used to implement the invention need not have orthogonally aligned points, as is the case for all of the constellations described thus far. For example, the constellation of FIG. 14 has radially-aligned points. There are eight supersymbols each of which is comprised of eight symbols. This constellation can thus support a two-dimensional coded modulation scheme with m = 2, k = 2. Each of the supersymbols can be identified based solely on angular information. Therefore, this constellation, like that of FIG. 8, allows for the use of coded modulation schemes which are expected to provide greater immunity against impulse noise for the more important bits than other constellations.

The foregoing merely illustrates the principles of the invention. For example, it is assumed in FIG. 1 that only one broadcast signal polarization is used. However, it is possible to double the bandwidth efficiency of the scheme by using a second set of coding circuitry to encode a second source-coded data stream in parallel with the first and to transmit the resulting coded modulated signal using a second polarization. Alternatively, a single data coding rail can be employed but its speed can be doubled by transmitting alternate signal points on the two polarizations.

It should be noted that, although in all the examples shown herein, the less important bits are always coded, this is not necessary. That is, uncoded bits can be used to select a symbol from the identified supersymbol. It should also be noted that, although in all of the examples shown herein, the minimum distance d_1 between superpoints is always greater than the minimum distance d_2 between points, this is not necessary. For example, d_1 can be equal to d_2 in FIG. 8. It should also be noted that, although in all the examples shown herein, only two classes of data elements are accommodated, the invention is not so limited. As many classes of data elements as desired can, in fact, be accommodated by dividing the class of less important bits into two or more subclasses and applying the principles of the invention to the coding of those subclasses in straightforward fashion. Moreover, although all the examples shown herein code either three or four information bits at a time, the invention is not in any way limited to these.

In the examples shown herein, encoders 114 and 115 are always of the same dimensionality. However, this is not necessary. For example, a two-dimensional code could be used for the more important data elements to identify a sequence of superpoints of a predetermined two-dimensional constellation. A four-dimensional code could then be used for the less important data elements to select points from sequential pairs of superpoints from that sequence. Conversely, a four-dimensional code could be used for the more important data elements and a two-dimensional code for the less important data elements.

In the examples shown herein, encoders 114 and 115 always implement 8-state trellis codes. However, this is not necessary. Codes having other than 8 states are equally usable. Moreover, other types of codes, such as block codes, can be used instead of trellis codes.

In some applications, it may be desired to provide for the possibility of phase rotations in the received signal caused by channel disturbance. In such applications, differential encoding circuitry may be included within channel encoder 114 to take care of this problem.

In addition, the invention is illustrated herein in the context of a digital video transmission system. However, it is equally applicable to other types of digital transmission systems. Moreover, although particular constellations are shown herein, numerous other constellations, which may be of any desired dimensionality, can be used.

It may also be noted that although the invention is illustrated herein as being implemented with discrete functional building blocks, e.g., source coders, scramblers, etc., the functions of any one or more of those building blocks can be carried out using one or more appropriate programmed processors, digital signal processing (DSP) chips, etc. Thus although each of the various "means" recited in the claims hereof may correspond, in some embodiments, to specific circuitry which is specifically designed to perform the function of just that means, it will be appreciated that such "means" may alternatively correspond, in other embodiments, to the combination of processor-based circuitry combined with stored program instructions which cause that circuitry to perform the function in question.

It will thus be appreciated that those skilled in the art will be able to devise numerous and various alternative arrangements which, although not explicitly shown or described herein, embody the principles of the invention and are within its spirit and scope.

Claims

1. A method

CHARACTERIZED BY the steps of

encoding a first group of data elements to generate a first expanded group of data elements,

identifying one of a plurality of supersymbols of a predetermined channel symbol constellation in response to the first expanded group of data elements, each supersymbol being comprised of a respective plurality of sym-

bols of the constellation,

selecting an individual one of the symbols of the identified supersymbol at least in response to a second group of data elements, and

applying to a communication channel a signal representing the selected symbol,

the minimum distance between at least ones of the symbols of at least one of the supersymbols being the same as the minimum distance betweeen the symbols of the constellation as a whole.

2. The invention of claim 1
CHARACTERIZED IN THAT
the selecting step includes the steps of
encoding said second group of data elements to generate a second expanded group of data elements, and
selecting said individual symbol in response to the second expanded group of data elements.

3. The invention of claim 1
CHARACTERIZED IN THAT
said encoding step includes the step of trellis coding said first group of data elements.

4. The invention of claim 1
CHARACTERIZED BY
the further step of generating said data elements by source coding input information in such away that said first group of data elements represents a portion of said information that is more important than the portion of said information represented by said second group of data elements.

5. The invention of claim 4
CHARACTERIZED IN THAT
said information is television information.

6. The invention of claim 1
CHARACTERIZED BY
the further step of rearranging the first expanded group of data elements prior to said identifying step.

7. The invention of claim 1
CHARACTERIZED IN THAT
at least one of said supersymbols is comprised of at least two non-contiguous groups of symbols.

8. A method for use in a receiver which receives intelligence communicated to said receiver by a transmitter, said transmitter being adapted to channel code successive groups of m+k data bits associated with respective symbol intervals via the steps, performed for each said interval, of a) encoding m of the bits of one of the groups using a first predetermined code to generate an expanded group of r bits, r > m; b) identifying a particular one of 2^r supersymbols of a predetermined channel symbol constellation as a function of the values of said r bits, each of said supersymbols being comprised of a plurality of symbols of said constellation assigned thereto and the minimum distance between the symbols of each supersymbol being the same as the minimum distance between the symbols of the constellation as a whole; c) generating a signal representing a selected one of the channel symbols of the identified one supersymbol, the selection being performed as a function of the values of the other k bits of said one group; and d) communicating said signal to said receiver over a communication channel;

said method CHARACTERIZED BY the steps of

receiving said signal from said channel, and

recovering said intelligence from the received signal, said recovering being carried out in response to information stored in said receiver about said first predetermined code, about said constellation, and the manner in which said symbols are assigned to their respective supersymbols.

9. The invention of claim 8
CHARACTERIZED IN THAT
the signal generating step in the transmitter includes the steps of a) encoding the other k bits of said one of the groups using a second predetermined code to generate a second expanded group of p bits, p > k; and b) selecting said individual symbol in response to the second expanded group of data bits,
and FURTHER CHARACTERIZED IN THAT
said recovering step is carried out further in response to information stored in said receiver about said second predetermined code.

10. The invention of claim 9
CHARACTERIZED IN THAT
said intelligence is television information.

11. The invention of claim 10
CHARACTERIZED IN THAT
said recovering step includes the step of decoding the received signal to recover said successive groups of data bits using maximum likelihood decoding.

12. Apparatus operative during each of a succession of symbol intervals for channel coding respective groups of m + k data bits, said apparatus

CHARACTERIZED BY

means for encoding m of the bits of one of the groups to generate an expanded group of r bits, r > m,

means for identifying a particular one of 2^r supersymbols of a predetermined channel symbol constellation as a function of the values of said r bits, each of said supersymbols being comprised of a plurality of symbols of said constellation, and

means for generating a signal representing a selected one of the channel symbols of the identified one supersymbol, the selection being performed as a function of the values of the other k bits of said one group,

the minimum distance between the symbols of each supersymbol being the same as the minimum distance between the symbols of the constellation as a whole.

13. The invention of claim 12

CHARACTERIZED IN THAT

said data bits represent television information.

14. The invention of claim 13

FURTHER CHARACTERIZED BY

means for generating said data bits by source coding input information in such a way that said m bits represent a portion of said information that is more important than the portion of said information represented by said k bits.

15. The invention of claim 14

CHARACTERIZED IN THAT

said generating means includes

means for encoding the other k bits of said one of the groups to generate an expanded group of p bits, p > k, and

means for selecting said individual symbol in response to said expanded group of p bits.

16. The invention of claim 15

CHARACTERIZED IN THAT

said m-bit and k-bit encoding means include means for trellis coding said m and k bits, respectively.

17. The invention of claim 15

CHARACTERIZED IN THAT

at least ones of said supersymbols are each comprised of at least two non-contiguous groups of symbols.

18. The invention of claim 16

FURTHER CHARACTERIZED BY

means for rearranging said expanded group of r bits prior to said identifying step.

19. An arrangement for use in a receiver which receives intelligence communicated to said receiver by a transmitter, said transmitter including apparatus for a) encoding a first stream of the data elements using a first predetermined code to generate a first expanded stream of data elements; b) identifying a sequence of supersymbols of a predetermined channel symbol constellation in response to the first expanded stream of data elements, the minimum distance between at least ones of the symbols of at least one of the supersymbols being the same as the minimum distance between the symbols of the constellation as a whole; c) encoding a second stream of data elements using a second predetermined code to generate a second expanded stream of data elements; d) selecting an individual one of the symbols of each supersymbol of said sequence at least as a function of the second expanded stream of data elements; and d) means for applying to a communication channel a signal representing the selected symbols,

said arrangement CHARACTERIZED BY

means for receiving the signal from the communication channel, and

means for carrying out a maximum likelihood decoding operation on the received signal to recover said first stream of data elements and for carrying out a second maximum likelihood decoding operation on the received signal to recover said second stream of data elements.

20. The invention of claim 19

CHARACTERIZED IN THAT

said first and second codes are trellis codes.

21. The invention of claim 19

CHARACTERIZED IN THAT

said intelligence is television information.

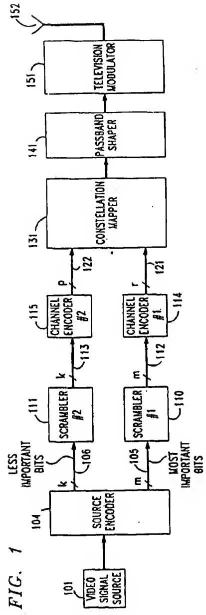

VIDEO SIGNAL SOURCE
101
SOURCE ENCODER
104
LESS IMPORTANT BITS
MOST IMPORTANT BITS
k
m
105
106
SCRAMBLER #2
111
SCRAMBLER #1
110
113
112
CHANNEL ENCODER #2
115
CHANNEL ENCODER #1
114
p
r
122
121
CONSTELLATION MAPPER
131
PASSBAND SHAPER
141
TELEVISION MODULATOR
151
152

FIG. 1

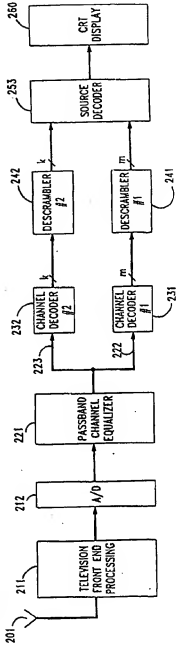

201
TELEVISION FRONT END PROCESSING
211
A/D
212
PASSBAND CHANNEL EQUALIZER
221
223
222
CHANNEL DECODER #2
232
CHANNEL DECODER #1
231
k
m
DESCRAMBLER #2
242
DESCRAMBLER #1
241
SOURCE DECODER
253
CRT DISPLAY
260

FIG. 2

FIG. 3

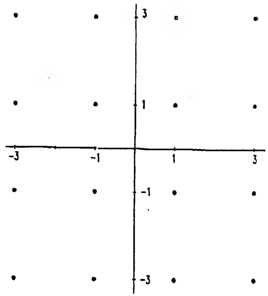


FIG. 4

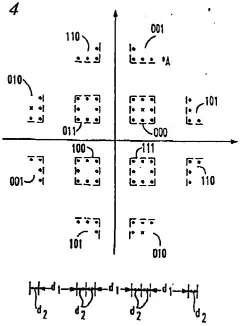


FIG. 5

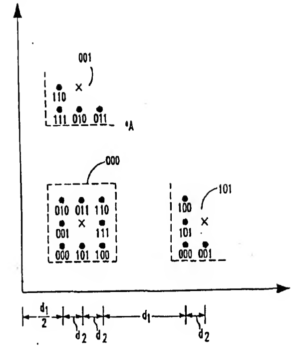

001
110
111 010 011
A
000
010 011 110
001 111
000 101 100
101
100
101
000 001
$\frac{d_1}{2}$
d_2 d_2
d_1
d_2

FIG. 6

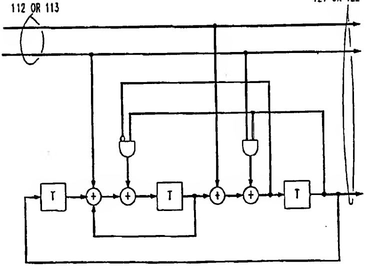

112 OR 113
121 OR 122
T
T
T

FIG. 7

TWO-DIMENSIONAL CONSTELLATION	d_1/d_2	FRACTION OF MOST IMPORTANT BITS (%)	CHANNEL ENCODERS	CODING GAIN FOR MOST IMPORTANT BITS (dB)	CODING GAIN FOR LESS IMPORTANT BITS (dB)
FIGURE 4	2.5	50	FIGURE 6	5.7	-2.8
FIGURE 4	3.5	50	FIGURE 6	6.6	-4.6
FIGURE 8	2	37.5	FIGURE 9	7.3	-0.1
FIGURE 8	3	37.5	FIGURE 9	9.1	-1.8
FIGURE 8	4	37.5	FIGURE 9	10.2	-3.2
FIGURE 12	2	62.5	FIGURE 13	5.1	-1.1
FIGURE 12	3	62.5	FIGURE 13	5.9	-3.5

FIG. 8

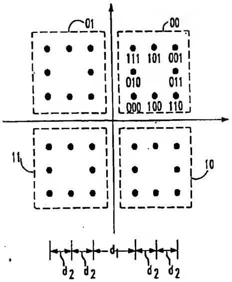

01
00
111 101 001
010 011
000 100 110
11
10
d1
d2 d2 d2 d2

FIG. 9

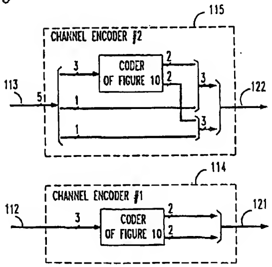

115
CHANNEL ENCODER #2
3
CODER OF FIGURE 10
2
2
3
3
113
5
1
1
122
114
CHANNEL ENCODER #1
112
3
CODER OF FIGURE 10
2
2
121

FIG. 10

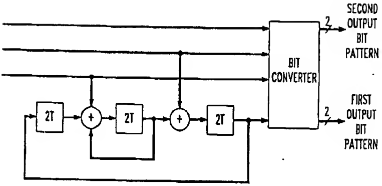


FIG. 11

INPUT BIT PATTERN OF BIT CONVERTER	FIRST OUTPUT BIT PATTERN	SECOND OUTPUT BIT PATTERN
0 0 0 0	0 0	0 0
0 0 0 1	0 0	0 1
0 0 1 0	0 0	1 1
0 0 1 1	0 0	1 0
0 1 0 0	0 1	0 1
0 1 0 1	0 1	1 1
0 1 1 0	0 1	1 0
0 1 1 1	0 1	0 0
1 0 0 0	1 1	1 1
1 0 0 1	1 1	1 0
1 0 1 0	1 1	0 0
1 0 1 1	1 1	0 1
1 1 0 0	1 0	1 0
1 1 0 1	1 0	0 0
1 1 1 0	1 0	0 1
1 1 1 1	1 0	1 1

FIG. 12

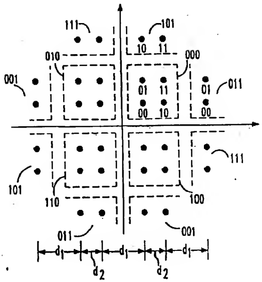

111
101
10
11
010
000
001
01
11
01
011
00
10
00
111
101
110
100
011
001
d1
d2
d1
d2
d1

FIG. 13

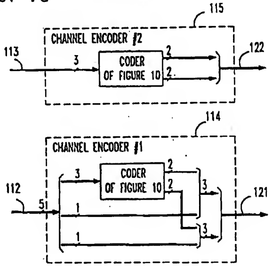

115
CHANNEL ENCODER #2
113
3
CODER
OF FIGURE 10
2
2
122
114
CHANNEL ENCODER #1
3
CODER
OF FIGURE 10
2
2
3
112
5
1
121
3
1

FIG. 14

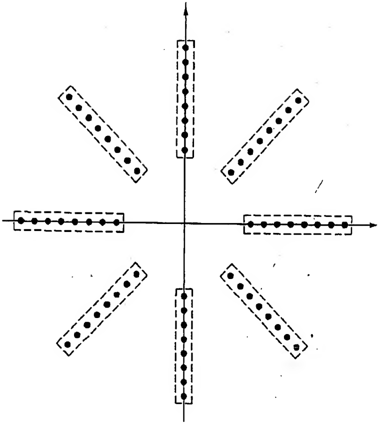
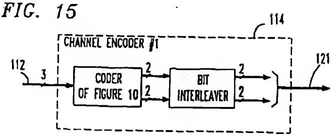


FIG. 15


114
CHANNEL ENCODER #1
112
3
CODER
OF FIGURE 10
2
2
BIT
INTERLEAVER
2
2
121

ES-93-TS (15)

H04L 5/00

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(54) Multi-subcarrier modulation for HDTV transmission.

(57) A high definition television (HDTV) signal is transmitted by a multi-subcarrier transmitter (100) in which each subcarrier (f1,f2...f12) is used to carry a different class of HDTV information.

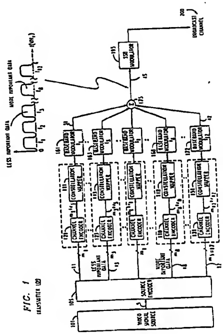

FIG. 1

EP 0 531 046 A2

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Background of the Invention

The present invention relates to the transmission of digital data, particularly the transmission of digital data that represents video signals.

It is generally acknowledged that some form of digital transmission will be required for the next generation of television (TV) technology, conventionally referred to as high definition television, or HDTV. This requirement is due mostly to the fact that much more powerful video compression schemes can be implemented with digital signal processing than with analog signal processing. However, in any HDTV digital transmission system, there are three major areas of concern that have to be addressed: graceful degradation, NTSC (National Television System Committee) co-channel interference and ghost cancellation.

A number of co-pending, commonly assigned United States patent applications disclose various techniques that provide graceful degradation in the reception quality at a TV set location. These are: V. B. Lawrence et al. entitled "Coding for Digital Transmission," Serial No. 07/611,225, filed on November 7, 1990; L.-F. Wei entitled "Coded Modulation with Unequal Error Protection," Serial No. 07/611,200, filed on November 7, 1990; J. D. Johnston et al. entitled "A High Definition Television Coding Arrangement with Graceful Degradation," Serial No. 07/625,349, filed on December 11, 1990; and H. Y. Chung et al. entitled "Multiplexed Coded Modulation with Unequal Error Protection," Serial No.07/627,156, filed on December 13, 1990. The Lawrence et al. patent application, for example, teaches the notion of characterizing the HDTV signal into classes of "more important" and "less important" information, which will then use a constellation of non-uniformly spaced signal points. This approach provides unequal error protection, i.e., more error protection for the more important information, and allows a graceful degradation in reception quality at the TV set location because, as the bit-error rate at the receiver begins to increase with increasing distance from the broadcast transmitter, it will be the bits that represent proportionately less of the TV signal information that will be the first affected.

However, although the above-mentioned patent applications teach advantageous techniques for providing unequal error protection to different classes of information, these approaches primarily address the problem of providing graceful degradation for an HDTV signal in a single carrier transmission environment and do not address the problems of NTSC co-channel interference and ghost cancellation.

NTSC co-channel interference is a result of the fact that any HDTV transmission scheme will co-exist with existing NTSC TV transmission schemes and will use the available NTSC frequency spectrum, or channel assignments. For example, in the New York City geographical area an HDTV television station may be assigned to broadcast on channel 3. However, there may also be an NTSC television station assigned to channel 3 in a neighboring geographical area such as Philadelphia. As a result, there will be parts of New Jersey that receive both the HDTV and NTSC television signals assigned to channel 3. This results in a geographical region of overlap of the NTSC and HDTV transmission signals in which the NTSC and HDTV signals interfere with each other. To reduce the interference from the HDTV signal to the existing NTSC signal, the transmitted power of the HDTV signal should be set at a value at least 10 dB below that of the NTSC signal so that the HDTV signal does not interfere with the NTSC signal. As a result, the HDTV signal is even more susceptible to interference from the NTSC signal. This NTSC interference must be reduced in order to ensure that the coverage area of the HDTV signal is large enough.

Finally, there is the problem of ghost cancellation. In any TV transmission scheme, reflection of the transmitted signal may occur that results in ghosting, which generally manifests itself in the form of double images. However, the problem of ghosting is compounded in an HDTV transmission scheme because of the use of compression algorithms to squeeze a full-bandwidth HDTV signal, e.g., 800 M bits/sec., into an NTSC 6 MHz channel. This necessitates the use of a complex equalizer to cancel the ghost images in an HDTV transmission scheme.

Before proceeding with a description of an illustrative embodiment, it should be noted that the various digital signaling concepts described herein—with the exception, of course, of the inventive concept itself—are all well known in, for example, the digital radio and voiceband data transmission (modem) arts and thus need not be described in detail herein. These include such concepts as multidimensional signaling using 2N-dimensional channel symbol constellations, where N is some integer; trellis coding; fractional coding; scrambling; passband shaping; equalization; Viterbi, or maximum-likelihood, decoding; etc.

Summary of the Invention

In accordance with the invention, a signal is divided into a plurality of classes of information which are encoded for different error protection levels. Each class of information is then modulated into a subchannel of the channel assigned to the signal. To further enhance signal reception, the subchannel assignments are based on noise and interference considerations.

The signal is separated into a plurality of classes of information such that at least one class of information is "more important" and is provided with more error protection than the remaining classes of information. The plurality of classes of information are then

frequency division multiplexed such that each class of information is modulated by a subcarrier into a subchannel within a frequency band.

In accordance with a feature of the invention, the effect of the NTSC co-channel interference is reduced by assigning the subchannel that carries the more important information to a frequency spectrum portion that is not subject to substantial NTSC interference. As a result, the more important data of the HDTV signal can still be recovered even in a fringe area where substantial NTSC co-channel interference is present.

In accordance with another feature of the invention, the use of multiple subcarrier results in longer symbol intervals and a flatter frequency response in each of the subchannels. As a result, a simpler equalizer can be used in the HDTV receiver to mitigate the effects of "ghosting."

Brief Description of the Drawing

FIG. 1 is a block diagram of a transmitter embodying the principles of the invention;

FIG. 2 is a block diagram of a receiver embodying the principles of the invention;

FIG. 3 is the frequency spectrum for an NTSC signal;

FIG. 4 is the frequency spectrum for an HDTV signal embodying the principles of the invention; and

FIGs 5 - 8 are illustrative 4, 8, 12 and 16 QAM signal constellations, respectively.

Detailed Description

In accordance with the principles of my invention, all three of the above-mentioned areas of concern in HDTV transmission are addressed. The signal is divided into a plurality of classes of information and each class of information is encoded to a different level of error protection. Each class of information is then modulated into a subchannel of the channel assigned to the signal.

Turning to FIG. 1, video signal source 101 generates an HDTV analog video signal representing picture information. As taught in the Lawrence et al. patent, this HDTV analog video signal is passed on to source encoder 105, which generates a digital signal comprising a plurality of "classes of information" in which at least one class of information is more important, i.e., contains "more important data," than the remainder of the classes of information that, therefore, contain "less important data." For example, the more important data represents that information, which is more important for reception of the information signal. In an HDTV signal, it is that information, which, if properly received, will form a rough picture, e.g., audio information, framing information, etc., and the less important data represents the information that comprises the remainder of the HDTV signal. As represented herein, source encoder 105 illustratively provides $k = 12$ classes of information with the class of information on lead 18 being "more important" than the other classes of "less important" information on the remaining leads, e. g., leads 11, 13 and 22. Illustratively, each class of information comprises a plurality of data bits, with an average of m_i bits, $1 \leq i \leq 12$, being generated on each lead for each signaling interval, which is of duration T seconds.

From FIG. 1 it can be seen that each class of information, which is represented by m_i bits, is processed by a channel encoder, a constellation mapper and a baseband modulator. For simplicity, the operation of transmitter 100 will be described, for the moment, in the context of the more important information on lead 18. A similar description would apply to the processing of each of the other classes of information. The more important information, which is represented by m_8 bits on lead 18, is input to channel encoder 128. The latter operates in accordance with known encoding techniques, such as trellis coding, and provides $m_8 + r_8$ data bits as output, where r_8 represents the average number of redundant bits introduced by channel encoder 128 in each signaling interval. (It should be noted that error correcting codes, such as a Reed-Solomon code, can also be used in place of, or in conjunction with, a coded modulation scheme.) The encoded output of channel encoder 128 is mapped, by constellation mapper 148, to a signal point, taken from a signal point constellation, in each signaling interval. It is assumed that the signal point constellation is representative of well-known uniformly-spaced QAM constellations such as is shown in FIGs. 5 to 8 for 4, 8, 12 and 16 signal point constellations.

Channel encoder 128 and constellation mapper 148, taken together, implement a particular coded modulation scheme that provides error protection to the more important class of information. The various coded modulation schemes that are implemented by the plurality of channel encoders, e.g., 121, 123, 128, 132, etc., and respective constellation mappers, 141, 143, 148, 152, etc., are chosen to provide unequal error protection to the plurality of classes of information such that the more important information is provided with more error protection. Unequal error protection can be implemented in a number of ways, such as different channel encoders, different constellations sizes and/or different symbol rates for the various channel encoders and constellation mappers. For example, referring to FIG. 1, all of the channel encoders can be identical. The signal constellation of constellation mapper 148 has, however, the smallest size compared to those of the other constellation mappers. For example, the constellation used by constellation mapper 148 is the 4-QAM of FIG. 5, while the 8-QAM, 12-QAM and 16-QAM of FIGs. 6-8 can be

used by the other constellation mappers. This assumes that the transmitted power for each subcarrier is the same, with the result that there is more separation between the signal points of the 4-QAM constellation of FIG. 5 (i.e., the spacing between the signal points), than in the constellations of FIGs. 6-8. Consequently, there is more error protection for the more important data, i.e., this provides unequal error protection for, and allows graceful degradation of, the HDTV signal.

Before proceeding, reference should be made to FIG. 3, which is a representative frequency spectrum for an illustrative NTSC analog TV baseband transmission signal that has a bandwidth of 6 MHz. (Although reference is made to the baseband signal, the actual transmitted signal is modulated to the respective frequency spectrum for a particular assigned channel. For example, channel 3 is transmitted in the frequency spectrum of 60 to 66 MHz.) In accordance with the invention, this 6 MHz NTSC bandwidth is divided into a number of subchannels, each subchannel assigned to one of a number of classes of information, which represent the HDTV signal. For the purposes of illustration, as shown in FIG. 4, the NTSC bandwidth is divided into 12 subchannels, with each subchannel having a bandwidth equal to 500 KHz, i.e., the NTSC bandwidth divided by the number of subchannels. Referring now back to FIG. 1, the HDTV signal is similarly divided into 12 classes of information. The output from each of the constellation mappers, e.g., 141, 143, 148, 152, etc. is provided to respective baseband modulators 161, 163, 168, 172, etc. The latter frequency modulates each of the encoded classes of information to a respective subcarrier, f_i (where $1 \leq i \leq 12$), such that each class of information is now provided in a respective subchannel. The outputs of the baseband modulators, e.g., 161, 163, 168, 172, etc., are summed, or frequency division multiplexed, by adder 175. The output of adder 175 is transmitted by single sideband (SSB) modulator 195. The latter is representative of conventional SSB modulation circuitry, e.g., oscillator, antenna, etc., and provides a broadcast HDTV signal to broadcast channel 200.

From FIG. 3, it can be seen that the energy of the NTSC transmission signal is generally concentrated in those frequency regions that contain the visual, chroma and aural carriers, at 1.25 MHz, 4.83 MHz and 5.75 MHz, respectively. As a result, any co-existing HDTV transmission signal in these frequency regions is subject to substantial interference. Therefore, and in accordance with a feature of the invention, the effect of NTSC co-channel interference can be reduced by assigning the more important information to a subchannel that is different from the subchannels that are subject to substantial interference from the NTSC visual, chroma and aural carriers. This is shown in FIG. 1, where the more important information is transmitted on subcarrier f_8, thereby avoiding the subchannels that are subject to substantial interference from the visual, chroma and aural carriers of the NTSC transmission signal, e.g., the subchannels associated with subcarriers f_3, f_{10}, f_{12}, etc. By avoiding those parts of the frequency spectrum of the NTSC transmission signal from which substantial interference is expected, the more important information is provided with more error protection than those classes of information that are assigned to those subchannels that overlap with the visual, chroma and aural carriers of the NTSC transmission signal. This additional error protection occurs even if all of the classes of information have the same encoding schemes. In addition, if an error occurs in those subchannels to which the less important information has been assigned, that information can simply be ignored by an HDTV receiver. For example, from FIG. 1, the less important information is assigned to subcarrier f_3, which is strongly interfered with by the visual carrier of the NTSC transmission signal. As a result, when an error occurs on this subchannel the less important information is ignored by the receiver. It should also be noted that those subchannels that experience substantial interference from the visual, chroma and aural carriers of the NTSC transmission signal can be intentionally left unused.

In accordance with another feature of the invention, the use of multiple subcarrier results in longer symbol intervals and a flatter frequency response in each of the subchannels. As a result, a simpler equalizer can be used in the HDTV receiver to mitigate the effects of ghosting. Further, a larger symbol interval provides more protection against noise spikes of short duration since fewer symbols would be effected.

Turning to the HDTV receiver of FIG. 2, the broadcast HDTV signal is received from broadcast channel 200 by receiver 300. The broadcast HDTV signal is received by SSB demodulator 395, which is representative of conventional reception and demodulation circuitry, e.g., the antenna, local oscillator, mixer, etc. SSB demodulator 350 provides a frequency multiplexed signal to each one of the plurality of bandpass filters, e.g., 341, 343, 348, 352, etc. For example, bandpass filter 348 filters out subcarrier f_8, which contains the more important information. This subcarrier is applied to equalizer 388 to compensate for intersymbol interference. The output of equalizer 388 is then provided to baseband demodulator 368, which provides a digital signal representing the received coded output to channel decoder 328. The latter decodes the received coded output to provide the more important data, on lead 68, to source decoder 305. Similarly, each of the other classes of information is decoded by receiver 300 through the respective demodulation and decoding circuitry. Source decoder 305 provides the inverse function of source en-

coder 105, of transmitter 100. Specifically, source decoder 305 takes into account the subchannel that each class of information is assigned to in a predetermined manner: for example, in order to recreate the analog HDTV signal, source decoder 305 knows a priori that the more important information is received on lead 68. As a result, source decoder 305 combines the various classes of information to provide the received analog HDTV signal to CRT display 301.

The foregoing merely illustrates the principles of the invention and it will thus be appreciated that those skilled in the art will be able to devise various alternative arrangements, which, although not explicitly described herein, embody the principles of the invention and are within its spirit and scope.

For example, as described hereinabove, all of the coded modulation schemes could be the same. Different symbol rates, or subchannels with different frequency bandwidths, could be used for the various classes of information. The use of a smaller symbol rate for the more important information would further mitigate the effects of ghosting, and hence provide more error protection for the more important data.

Also, it should be observed that one subchannel can be used to carry other information in addition to the plurality of classes of information of the HDTV signal. For example, a subchannel with a fixed coding and modulation format, which carries the more important information, can be used to transmit information as to the coding and modulation formats used on the other subchannels so that, illustratively, a variable bit rate can be used for each of class of information.

In addition, more than one class of information may be carried by each subchannel and a nonuniformly-spaced signal point constellation can also be used. Alternatively, more than one constellation may be used by each subchannel, each constellation being for one class of information and this constellation being time-division-multiplexed.

It may also be noted that the number of subcarriers used is not restricted to twelve but can be any number greater than one. Further, the implementation of the frequency division multiplexed scheme can include overlapping of the spectra of different subcarriers and/or different modulation schemes. Also, other communications system components can be used as well, such as an interleaver to protect against bursty noise. In addition, although the invention is illustrated herein as being implemented with discrete functional building blocks, e.g., trellis encoders, constellation mappers, etc., the functions of any one or more of those building blocks can be carried out using one or more appropriate programmed processors, digital signal processing (DSP) chips, etc.

Claims

1. A method for processing an information signal comprising the steps of

   separaring the information signal into a plurality of classes of information such that at least one of the plurality of classes of information is more important for reception of the information signal than the other ones of the plurality of classes of information,

   encoding the plurality of classes of information to provide a plurality of encoded symbols such that the more important information has more error protection than the remaining ones of the plurality of classes of information, and

   modulating the plurality of encoded symbols into a plurality of subchannels within a frequency band, each subchannel occupying a different frequency spectrum.

2. The method of claim 1 wherein the frequency band is assigned to a different signal.

3. The method of claim 2 wherein the modulating step places the more important information into a subchannel that does not contain a carrier of the different signal.

4. The method of claim 1 wherein the encoding step includes the steps of

   channel encoding each one of the plurality of classes of information to provide a plurality of encoded outputs, and

   mapping each one of the plurality of encoded outputs to a signal constellation to provide the plurality of encoded symbols.

5. The method of claim 4 wherein the channel encoding step for the more important information is different from the channel encoding step of at least one other of the plurality of classes of information.

6. The method of claim 5 wherein the channel encoding step operates in accordance with an error correcting code.

7. The method of claim 5 wherein the channel encoding step operates in accordance with coded modulation.

8. The method of claim 5 wherein the channel encoding step operates in accordance with coded modulation and an error correcting code.

9. The method of claim 4 wherein the mapping step for the more important information uses a signal point constellation that is different from at least

one other of the signal point constellations used for the other plurality of classes of information.

10. The method of claim 1 wherein the encoding step uses a symbol rate for the more important class of information that is different from the symbol rate of a least one other class of information.

11. Apparatus for processing an information signal comprising

means for separating the information signal into a plurality of classes of information such that at least one of the plurality of classes of information is more important than the other ones of the plurality of classes of information for reception of the information signal,

means for providing unequal levels of error protection to the plurality of classes of information to provide a plurality of encoded symbols for each one of the plurality of classes of information such that the more important information has more error protection than the remaining ones of the plurality of classes of information, and

means for modulating the plurality of encoded symbols for each one of the plurality of classes of information to a plurality of subchannels within a frequency band, each subchannel occupying a different frequency spectrum.

12. The apparatus of claim 11 wherein the frequency band is assigned to a different signal.

13. The apparatus of claim 12 wherein the means for modulating places the more important information into a subchannel that does not contain a carrier of the different signal.

14. The apparatus of claim 11 wherein the means for providing unequal error protection further comprises

channel encoding means for each one of the plurality of classes of information to provide a plurality of encoded outputs, and

means for mapping each one of the plurality of encoded outputs to a signal constellation to provide the plurality of encoded symbols.

15. The apparatus of claim 14 wherein the channel encoding means for the more important information is different from the channel encoding means of at least one other of the plurality of classes of information.

16. The apparatus of claim 14 wherein the means for mapping the more important information uses a signal point constellation that is different from at least one other of the signal point constellations used for the other plurality of classes of information.

17. The apparatus of claim 11 wherein the means for providing unequal error protection uses a symbol rate for the more important class of information that is different from the symbol rate of at least one other class of information.

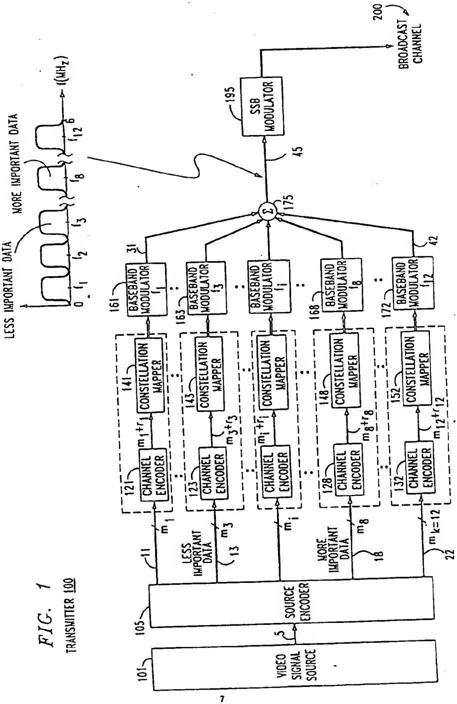
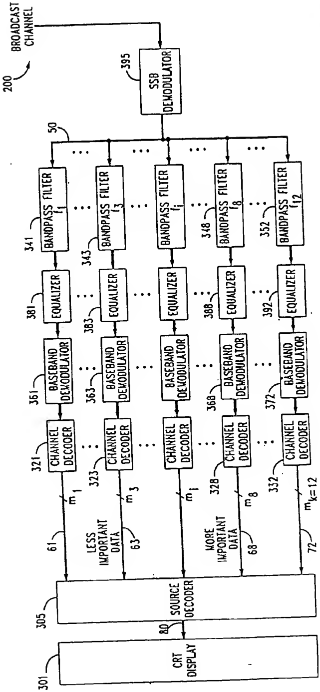


FIG. 1

TRANSMITTER 100

FIG. 2 RECEIVER 300



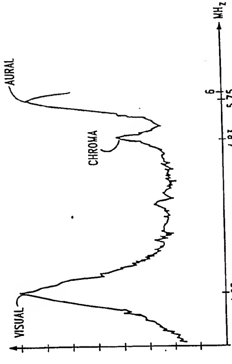

AURAL
VISUAL
CHROMA
10dB/DIV
0
1.25
4.83
5.75
6
MHz

FIG. 3

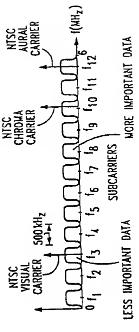
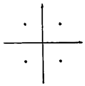

NTSC VISUAL CARRIER
500 kHz
NTSC CHROMA CARRIER
NTSC AURAL CARRIER
0
f1 f2 f3 f4 f5 f6 f7 f8 f9 f10 f11 f12
6
f(MHz)
LESS IMPORTANT DATA
SUBCARRIERS
MORE IMPORTANT DATA

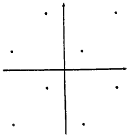
FIG. 4

FIG. 5



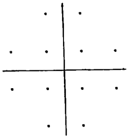
4 QAM SIGNAL CONSTELLATION

FIG. 6



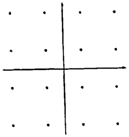
8 QAM SIGNAL CONSTELLATION

FIG. 7



12 QAM SIGNAL CONSTELLATION

FIG. 8



16 QAM SIGNAL CONSTELLATION

Office européen des brevets

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(54) Multi-subcarrier modulation for HDTV transmission.

(57) The HDTV signal is divided into a plurality of information classes, which are each encoded with an individual level of error correction ; and which are each transmitted on an individual subcarrier. The subchannel assignments depend upon noise and interference considerations.

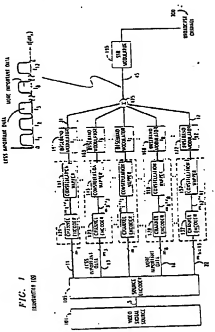

FIG. 1

EP 0 531 046 A3

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European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 92 30 7802

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.5)
X	RUNDFUNKTECHNISCHE MITTEILUNGEN vol. 35, no. 2, March 1991, NORDERSTEDT DE pages 45 - 66 PLENGE 'DAB - Ein neues Hörfunksystem. Stand der Entwicklung und Wege zu seiner Einführung'	1,10,11, 17	H04L5/06 H04N7/13 H04N7/08
Y	* abstract *	2-6, 12-15	
A	* page 54, paragraph 5.3 - page 59, line 53; figure 5 *	7,8,10	

Y	US-A-4 884 139 (POMMIER) * column 2, line 3 - line 40 * * column 4, line 9 - line 27; claim 1; figure 1 *	2,3,12, 13	

Y	WO-A-8 607 223 (TELEBIT CORP.) * abstract *	4-6,14, 15	
A	* page 4, line 25 - page 5, line 9 * * page 11, line 17 - line 24 * * page 13, line 25 - line 36 * * page 18, line 21 - line 25; figures 1-3,5 *	2,3,9,16	TECHNICAL FIELDS SEARCHED (Int. Cl.5) H04L H04N

P,X	EP-A-0 448 492 (ETAT FRANCAIS) * abstract * * page 4, line 5 - line 30 * * page 6, line 45 - page 7, line 29 * * claims 1,4; figure 3 *	1-17	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	20 APRIL 1993	WAGNER U.

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
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訂正有り

⑲日本国特許庁

⑪特許出願公開

公開特許公報

昭53—137657

⑤Int. Cl.² H 03 D 3/00 H 04 L 27/22　識別記号　⑫日本分類 98(5) E 22　庁内整理番号 6628—53　④公開 昭和53年(1978)12月1日

発明の数 1
審査請求 未請求

（全 4 頁）

⑭位相復調装置

㉑特 願 昭52—52192
㉒出 願 昭52(1977)5月7日
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明細書

1. 発明の名称

位相復調装置

2. 特許請求の範囲

バーストモードで伝送され、そのプリアンブル部のユニークワードが2相PSK波で、かつデータ部が4相PSK波で形成されたバーストモードPSK波信号の搬送波を再生する2相用搬送波再生器および4相用搬送波再生器、この2相用搬送波再生器の出力を基準信号として上記2相PSK波を同期検波し、上記ユニークワードを検出するユニークワード検出部、上記2相用搬送波再生器の出力と上記4相用搬送波再生器の出力および上記2相用搬送波再生器の出力と$\pi/2$ (rad) 移相された上記4相用搬送波再生器の出力をそれぞれ位相比較してその位相差に応じて2値符号化すると共に、これらの符号出力値を上記ユニークワード検出部の検出値に応じて直接あるいは反転させて出力する位相比較器を備え、上記位相比較器の出力値により、上記4相用搬送波再生器の出力を基準信号として同期検波される4相P.S Kの復調信号（データ）の位相不確定性を除去するようにしたことを特徴とする位相復調装置。

3. 発明の詳細な説明

この発明はバーストモードPSK波信号を復調する位相復調装置に関するもので、特にその復調時において生ずる位相不確定性 (phase ambiguity) の改善に関するものである。

従来のこの種の装置は第1図に示すように、バーストモードPSK波信号（この場合、プリアンブル部のユニークワードおよびデータ部は共に4相PSK波）は入力端子(1)に入力され、この入力端子を介して4相用搬送波再生器(2)、4相用の位相検波器(3)，(4)およびビットタイミング再生器(6)にそれぞれ入力されており、4相用搬送波再生器(2)ではその搬送波（無変調波）が、ビットタイミング再生器(6)ではビットタイミング波が再生出力される。この場合、4相用搬送波再生器(2)から出力される搬送波の位相は第2図に示すように0°，90°

，180°，－90°の4状態のうちいずれかをとり、その出力位相はいずれの状態になるかは不確定である。

位相検波器(3)および(4)はこの4相用搬送波再生器(2)の出力およびこれをπ/2移相する移相器(6)の出力を基準信号として、入力端子(1)から入力される4相PSK波信号を同期検波し、その基底帯(baseband)信号をそれぞれ出力する。これらの基底帯域信号は識別再生器(7)および(8)にそれぞれ入力され、識別再生器(7)および(8)はビットタイミング再生器(5)から出力されるビットタイミング波によりこれらの基底帯域信号を各ビット毎に波形整形して復調信号を得、これをユニークワード検出器(9)およびアンビギュイティスイッチ(10)にそれぞれ入力する。

ここで、識別再生器(7)および(8)で得られる復調信号は上述した位相不確定性を有しており、4相用搬送波再生器(2)の出力位相が0°以外の時は誤つた復調信号を得ていることになる。そこで、入力端子(1)に入力されるバーストモードPSK波信号には、この位相不確定性を除去し、バーストタイミングを得るためのユニークワード(以下UWと云う)が各バースト毎に挿入されており、このUWが4相PSK波で伝送される伝送系では、互いに直交する2つのUW(P・Q)が送信されている。

ユニークワード検出器(9)に入力された復調信号すなわち復調されたUWはその復調時における位相不確定性により(P・Q)、($\bar{Q}$・P)、(Q・$\bar{P}$)、($\bar{P}$・$\bar{Q}$)の4状態をとり、いずれかの状態がユニークワード検出器(9)により検出され、その検出値がアンビギュイティ制御器(11)に入力される。アンビギュイティ制御器(11)はその検出値の位相状態を判別し、位相ずれに応じた制御信号をアンビギュイティスイッチ(10)に与える。このアンビギュイティスイッチ(10)はこの制御信号により識別再生器(7)および(8)から出力される復調信号の位相不確定性を除去して出力端子(12a)および(12b)に出力する。

以上述べた従来の装置においては、入力端子(1)に入力されるバーストモードPSK波信号の搬送波信号電力対雑音電力比(以下CNRと云う)が良い場合(符号誤り率(以下BERと云う)が10^{-4}以下に相当)は問題ないが、悪い場合(BERが10^{-4}以上に相当)はユニークワード検出器(9)が誤動作し、UWの検出を誤ることがある。

しかるに最近ではCNRが悪い場合、例えばBERが10^{-2}以下において誤動作があつてはならない(UWの検出誤りが10^{-8}以下で、しかも位相不確定性が除去されていること)と云うような要求があり、この要求を満足させるには、従来の装置では不可能であつた。

この発明はこのような点にかんがみてなされたもので、CNRが悪化しても誤動作することなく確実に復調できる位相復調装置を提供するものである。

以下第3図に示すこの発明の一実施例について説明する。第3図において、(20)は2相用搬送波再生器、(23)は2相用の位相検出器(22)、識別再生器(24)およびユニークワード検出器(25)で構成されたユニークワード検出部、(26)は位相比較器、(27)はアンビギュイティ制御器である。なお、(1)～(8)，(10)，(12a)，(12b)は第1図の従来装置と同一であるので説明は省略する。

このような構成において、入力端子(1)には第4図に示すようなプリアンブル部(ユニークワード)が2相PSK波で、データ部が4相PSK波からなるバーストモードPSK波信号が入力されるものとする。このバーストモードPSK波信号は4相用搬送波再生器(2)、2相用搬送波再生器(20)にそれぞれ入力され、その搬送波に再生されるのであるが、この場合、4相用搬送波再生器(2)の出力は上述したように4状態の位相不確定性をもち、また2相用搬送波再生器(20)は2相用のためその出力は2状態の位相不確定をもつ(この位相状態を45°，225°とする)。

すなわち、4相用搬送波再生器(2)の出力をa_1、移相器(6)の出力をa_2、また2相用搬送波再生器(20)の出力をa_3とすると、

$$a_1 = \sin\left\{\omega_0 t + \frac{n\pi}{2}\right\} \qquad (1)$$

$$a_2 = \sin\left\{\omega_0 t + \frac{\pi}{2} + \frac{n\pi}{2}\right\} \qquad (2)$$

$$a_3 = \sin\left\{\omega_0 t + \frac{\pi}{4} + m\pi\right\} \qquad (3)$$

となる。なお、nは4相用搬送波再生器(2)の位相不確定性を表わし、n＝0（0°の場合）、1（90°）、2（180°）、3（−90°）。またmは2相用搬送波再生器(20)の位相不確定性を表わし、m＝0（45°）、1（225°）とする。

この2相用搬送波再生器(20)の出力a_3を基準信号として、入力端子(1)から入力されるプリアンブル部の2相PSK波は位相検波器(23)において同期検波され、その検波出力は識別再生器(24)でビットタイミング再生器(6)から出力されるビットタイミング波により波形整形され、復調されたUWを出力する。このUWはRあるいは$\bar{R}$の値を有し、このUW値はユニークワード検出器(25)により検出される。この場合Rを検出すれば2相用搬送波再生器(20)の出力位相は45°、$\bar{R}$を検出すれば225°である。

ところで、位相検波器(20)は2相用であるので、4相用の位相検波器(3)および(4)に比べ、その検波出力レベルは3dB高くなり、このことはプリアンブル部のユニークワードが4相PSK波の場合BERが10^{-2}に相当するのに対し、2相PSK波の場合BERが4×10^{-4}に相当する。また2相用搬送波再生器(20)の方が4相用搬送波再生器(2)に比べ、その出力の雑音電力は減少する。したがつて、ユニークワード検出器(25)は従来装置におけるユニークワード検出器(9)に比べ、UWの検出を誤る確率は少なくなる。

位相比較器(26)は2相用搬送波再生器(20)の出力a_3と4相用搬送波再生器(2)の出力a_1および移相器(6)の出力a_2をそれぞれ同期検波する。この検波出力の直流分をA_1およびA_2とすると、

$$A_1 = \frac{1}{\sqrt{2}}\cos\left(\frac{2n-1}{4}\pi - m\pi\right) + \frac{1}{2} \qquad (4)$$

$$A_2 = \frac{1}{\sqrt{2}}\cos\left(\frac{2n+1}{4}\pi - m\pi\right) + \frac{1}{2} \qquad (5)$$

となる。すなわち

$$\left.\begin{array}{llr}
m=0 & n=0\text{の場合} & (A_1,\ A_2) = (1.1) \\
 & 1 & (1.0) \\
 & 2 & (0.0) \\
 & 3 & (0.1) \\
m=1 & n=0\text{の場合} & (A_1,\ A_2) = (0.0) \\
 & 1 & (0.1) \\
 & 2 & (1.1) \\
 & 3 & (1.0)
\end{array}\right\} \quad (6)$$

となる。

ユニークワード検出器(25)はm＝0の場合Rを、m＝1の場合は$\bar{R}$を検出するので、これを位相比較器(26)に与え、$\bar{R}$を検出した時のみ位相比較器(26)の出力（A_1、A_2）の符号を反転させれば(6)式の符号値はmの値にかかわらず

$$\left.\begin{array}{lr}
n=0\text{の場合} & (A_1,\ A_2) = (1.1) \\
1 & (1.0) \\
2 & (0.0) \\
3 & (0.1)
\end{array}\right\} \quad (7)$$

となる。

この出力（A_1，A_2）をアンビギュイティ制御器(27)に入力して位相状態を判別し、位相ずれに応じた制御信号がアンビギュイティスイッチ(10)に与えられる。アンビギュイティスイッチ(10)はこの制御信号により、識別再生器(7)および(8)から出力される4相PSK波の復調信号（データ）の位相不確定性を除去して出力端子（12a）および（12b）に出力する。

以上はTDMA 4相PSK波のバーストモードの伝送系について説明したが、この発明はこれに限らず4相のSCPC-PSKに使用してもよい。 3字挿入

以上のようにこの発明に係る位相復調装置では受信CNRが悪くても誤動作をおこしにくいものであるから、例えば、アンテナを小形化してアンテナ利得を下げたり、低雑音増幅器の雑音温度を上げるなどによつて衛星通信システムや地上通信システムの低コスト化ができる利点がある。

4. 図面の簡単な説明

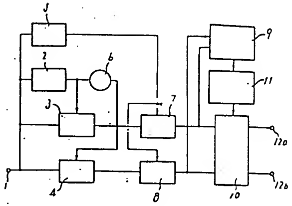
第1図は従来の位相復調装置の回路構成を示す系統図、第2図は第1図の動作を説明するための説明図、第3図はこの発明の一実施例の回路構成を示す系統図、第4図は第3図の説明図である。

図中、(21)は4相用搬送波再生器、(22)は2相用搬送波再生器、(23)はユニークワード検出部、(24)は位相比較器、(25)はアンビギュイティ制御器である。

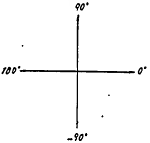
なお図中、同一あるいは相当部分には同一符号を付して示してある。

代理人　葛野信一

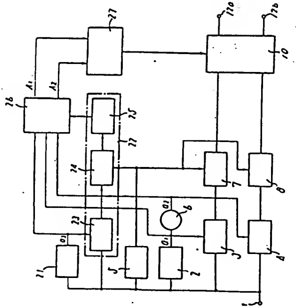
第1図



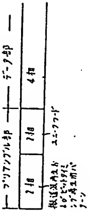
第2図



第3図



第4図



特許法第17条の2の規定による補正の掲載

昭和52年特許願第52192号(特開昭53-137657号 昭和53年12月1日発行 公開特許公報 53-1377号掲載)については特許法第17条の2の規定による補正があったので下記のとおり掲載する。 7(3)

Int.Cl³.	識別記号	庁内整理番号
H03D 3/00		7402-5J
H04L 27/22		7240-5K

58- 1078

手続補正書(自発)

昭和58年3月31日

特許庁長官殿

1. 事件の表示 特願昭52-052192号

2. 発明の名称
位相復調装置

3. 補正をする者

事件との関係 特許出願人
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(連絡先03(213)3421特許部)

(1)

5. 補正の対象
明細書の発明の詳細な説明および図面の簡単な説明の欄

6. 補正の内容

(1) 明細書第3頁第14行～第15行，第4頁第16行，および第4頁第16行，第10頁第8行，および第10頁第4行にそれぞれ「アンビギユテイスイツチ(10)」とあるのを「アンビギユイテイスイツチ(10)」と訂正する。

(2) 同第4頁第12行に「アンビユユテイ」とあるのを「アンビギユイテイ」と訂正する。

(3) 同第4頁第18行，第6頁第8行，および第10頁第2行にそれぞれ「アンビギユテイ」とあるのを「アンビギユイテイ」と訂正する。

(4) 同第4頁第5行に「(P・Q)」とあるのを「(P，Q)」と訂正する。

(5) 同第4頁第9行～第10行に「(P・Q)、($\overline{Q}$・P)、(Q・$\overline{P}$)、($\overline{P}$・$\overline{Q}$)」とあるのを「(P，Q)、($\overline{Q}$，P)、(Q，$\overline{P}$)、($\overline{P}$，$\overline{Q}$)」と訂正する。

(6) 同第5頁第4行に「(10^{-4}」とあるのを「10^{-4}」と訂正する。

(7) 同第11頁第7行に「アンビギユテイ」とあるのを「アンビギユイテイ」と訂正する。

以上

P 1[illegible]/P 11166/P 11667

No. 53-137657

SPECIFICATION

1. Title of the Invention

Phase demodulating apparatus

2. What is claimed is:

A phase demodulating apparatus comprising a two-phase carrier regenerator and a four-phase carrier regenerator for regenerating carries of burst mode PSK wave signals, transmitted in burst mode, of which unique word of preamble unit is formed of two-phase PSK wave, and data unit is formed of four-phase PSK wave, a unique word detector for detecting said unique word by synchronously detecting the two-phase PSK wave by using the output of the two-phase carrier regenerator as reference signal, and a phase comparator for comparing the phases of the output of said two-phase carrier regenerator, the output of said four-phase carrier regenerator, and the output of said four-phase carrier regenerator shifted in phase by π /2 (rad) from the output of said two-phase carrier regenerator, coding in 2-level value depending on each phase difference, and issuing these code output values directly or by inverting depending on the detected value of the unique word detector, wherein phase ambiguity of demodulated signal (data) of said four-phase PSK detected synchronously is removed, using the output of said four-phase carrier regenerator as the reference signal, by the output value of said phase comparator.

3. Detailed Description of the Invention

The present invention relates to a phase demodulating apparatus for demodulating burst most PSK wave signal, and more particularly to an improvement of phase ambiguity occurring at the time of demodulation.

Hitherto, the apparatus of this kind was constructed as shown in Fig. 1, in which a burst mode PSK wave signal (in this case, the unique word of preamble unit and data unit are both four-phase PSK waves) is put into an input terminal (1), and further through this input terminal, it is fed into a four-phase carrier regenerator (2), four-phase detectors (3), (4), and a bit timing regenerator (5), and the four-phase carrier regenerator (2) issues its carrier (non-modulated wave), and the bit timing regenerator (5) issues a bit timing wave. In this case, the phase of the carrier issued from the four-phase carrier regenerator (2) is any one of four states, that is, 0 °, 90 °, 180 °, and -90 °, as shown in Fig. 2, and it is ambiguous in which state the output phase settles.

The phase detectors (3) and (4) synchronously detect the four-phase PSK wave signals entered from the input terminal (1) on the basis of the reference signal of the output of the four-phase carrier regenerator (2) and phase shifter (6) for shifting its phase by $\pi/2$, and issue their baseband signals, respectively. These baseband signals are fed respectively into discriminative regenerators (7) and (8), and the discriminative regenerators (7) and (8) shape the waveforms of

these baseband signals in every bit by the bit timing wave issued from the bit timing regenerator (5), and obtain demodulated signals, then feed them into a unique word detector (9) and an ambiguity switch (10).

The demodulated signals obtained in the discriminative regenerators (7) and (8) involve the phase ambiguity mentioned above, and unless the output phase of the four-phase carrier regenerator (2) is 0 °, wrong demodulated signal is obtained. Accordingly, in the burst mode PSK wave signal entered in the input terminal (1), a unique word (hereinafter called UW) is inserted in every burst for obtaining the burst timing, and in the transmission system for transmitting this UW in four-phase PSK wave, mutually orthogonal two UW (P, Q) are transmitted.

The demodulated signal fed into the unique word detector (9), that is, the demodulated UW may exist in one of four states (P, Q), ($\overline{Q}$, P), (Q, $\overline{P}$), and ($\overline{P}$, $\overline{Q}$), depending on the phase ambiguity at the time of demodulation, and any one state is detected by the unique word detector (9), and the detected value is put into an ambiguity controller (11). The ambiguity controller (11) judges the phase state of the detected value, and gives a control signal depending on the phase deviation to an ambiguity switch (10). The ambiguity switch (10) removes the phase ambiguity of the modulated signals issued from the discriminative regenerators ((7) and (8) by this control signal, and issues to output terminals (12a) and (12b).

In the conventional apparatus described so far, as far as the ratio of the carrier signal electric power to the noise

electric power (hereinafter called CNR) of the burst mode PSK wave signal entered in the input terminal (1) is favorable (the bit error rate (BER) corresponding to 10^{-4} or less), there is no problem, but inferior (BER corresponding to over 10^{-4}), the unique word detector (9) may malfunction, and detection of UW may fail.

Recently, therefore, when the CNR is poor, for example, it is required that no malfunction should occur at the BER of less than 10^{-2} (that is, the detection error of UW be 10^{-8} or less, and phase ambiguity should be removed), this requirement could not be satisfied by the conventional apparatus.

The invention is devised in the light of such background, and it is hence an object thereof to present a phase demodulating apparatus capable of demodulating securely without malfunctioning even if the CNR is worsened.

An embodiment of the invention shown in Fig. 3 is described. In Fig. 3, reference numeral (21) is a two-phase carrier regenerator, (22) is a two-phase detector, (23) is a unique word detector composed of discriminative regenerator (24) and unique word detector (25), (26) is a phase comparator, and (27) is an ambiguity controller. Reference numerals (1) to (8), (10), (12a), and (12b) are same as in the conventional apparatus in Fig. 1, and their description is omitted.

In this constitution, suppose the input terminal (1) has received the burst mode PSK wave signal composed of two-phase PSK wave in the preamble unit (unique word) and four-phase PSK wave in the data unit as shown in Fig. 4. This burst mode PSK

wave signal is put into the four-phase carrier regenerator (2) and two-phase carrier regenerator (21), and regenerated into carriers, and in this case it is supposed that the output of the four-phase carrier regenerator (2) has four states of phase ambiguity as mentioned above, and that the output of the two-phase carrier regenerator (21) has two states of phase ambiguity for the sake of two phases (these phase states are 45 ° and 225 °).

That is, supposing the output of the four-phase carrier regenerator (2) to be a_1, the output of the phase shifter (6) to be a_2, and the output of the twoOphaes carrier regenerator (21) to be a_3,

$$a_1 = \sin\{\omega_0 t + \frac{n\pi}{2}\} \qquad (1)$$

$$a_2 = \sin\{\omega_0 t + \frac{\pi}{2} + \frac{n\pi}{2}\} \qquad (2)$$

$$a_3 = \sin\{\omega_0 t + \frac{\pi}{4} + m\pi\} \qquad (3)$$

are obtained. Herein, n denotes the phase ambiguity of the four-phase carrier regenerator (2), being n = 0 (in the case of 0 °), 1 (90 °), 2 (180 °), and 3 (-90 °), and m denotes the phase ambiguity of two-phase carrier regenerator (21), being m = 0 (45 °), 1 (225 °).

Using the output a_3 of the two-phase carrier regenerator (21) as the reference signal, the two-phase PSK wave of the preamble unit entered from the input terminal (1) is synchronously detected by the phase detector (23), its detection output is shaped in waveform by the bit timing wave issued from the bit timing regenerator (5) by the discriminative

regenerator (24), and the demodulated UW is issued. This UW has a value of R or $\overline{R}$, and this UW value is detected by the unique word detector (25). In this case, when detecting R, the output phase of the two-phase carrier regenerator (21) is 45 °, and when detecting R-, it is 225 °

Incidentally, since the phase detector (21) is for two phases, and as compared with the four-phase detectors (3) and (4), its detection output level is higher by 8 dB, that is, when the unique word of the preamble unit is four-phase PSK wave, the BER corresponds to 10^{-2}, or in the case of two-phase PSK wave, the BER corresponds to 4 x 10^{-4}. Besides, the two-phase carrier regenerator (21) decreases in the noise power of its output as compared with the two-phase carrier regenerator (2). Therefore, the unique word detector (25) is lower in the probability of detection error of UW as compared with the unique word detector (9) in the prior art.

The phase comparator (26) synchronously detects the output a3 of the two-phase carrier regenerator (21), the output a1 of the four-phase carrier regenerator (2), and the output a3 of the phase shifter (6). The DC components of the detection output A_1 and A_2 are

$$A_1 = \frac{1}{\sqrt{2}}\cos(\frac{2n-1}{4}\pi - m\pi)+\frac{1}{2} \quad (4)$$

$$A_2 = \frac{1}{\sqrt{2}}\cos(\frac{2n+1}{4}\pi - m\pi)+\frac{1}{2} \quad (5)$$

That is,

In the case of m=0, n=0, $(A_1, A_2 = (1.\ 1)$ (6)

In the case of m=0, n=1, $(A_1, A_2 = (1.\ 0)$

In the case of m=0, n=2, $(A_1,A_2 = (0.\ 0)$

In the case of m=0, n=3, $(A_1,A_2 = (0.\ 1)$

In the case of m=1, n=0, $(A_1,A_2 = (0.\ 0)$

In the case of m=1, n=1, $(A_1,A_2 = (0.\ 1)$

In the case of m=1, n=2, $(A_1,A_2 = (1.\ 1)$

In the case of m=1, n=3, $(A_1,A_2 = (1.\ 0)$

Since the unique word detector (26) detects R in the case of m=0, and detector $\overline{R}$ in the case of m=1, by giving it to the phase comparator (26), the code of the output (A_1, A_3) of the phase comparator (26) is inverted only when R- is detected, the value of formula (6)is as follows regardless of the value of m:

In the case of n=0, $(A_1,A_2) = (1.\ 1)$ (7)

In the case of n=1, $(A_1,A_2) = (1.\ 0)$

In the case of n=2, $(A_1,A_2) = (0.\ 0)$

In the case of n=3, $(A_1,A_2) = (0.\ 1)$

Feeding this output (A_1, A_3) into the ambiguity controller (27), the phase state is judged, and the control signal depending on the phase deviation is given to the ambiguity switch (10). By this control signal, the ambiguity switch (10) removes the phase ambiguity of demodulated signal (data) of four-phase PSK wave issued from the discriminative regenerators (7) and (8), and issues to the output terminals (12a) and (12b).

So far is explained about the transmission system of the burst mode of the TDMA four-phase PSK wave burst mode, but not limited to this, the invention may be applied also in the SCPC-PSK.

Thus, in the phase demodulating apparatus of the invention, malfunction hardly occurs if the reception CNR is poor, and therefore, the antenna gain may be lowered by reducing the size of antenna, or the noise temperature of the low noise amplifier may be raised, so that the satellite communication system or ground communication system may be lower in cost.

4. Brief Description of the Drawings

Fig. 1 is a block diagram showing a circuit configuration of a conventional phase demodulating circuit, Fig. 2 is an explanatory diagram for explaining the operation of Fig. 1, Fig. 3 is a block diagram showing a circuit configuration of an embodiment of the invention, and Fig. 4 is an explanatory diagram of Fig. 3.

In the drawings, reference numeral (2) is a four-phase carrier regenerator, (21) is a two-phase carrier regenerator, (22) is a unique word detector, (26) is a phase comparator, and (27) is an ambiguity detector.

Same parts or corresponding parts in the drawings are identified with same reference numerals.

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Fig. 4

Preamble unit

Data unit

2 phases

2 phases

4 phases

Pattern for regeneration of carrier and regeneration of bit timing

Unique word

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(54) Method and apparatus for combining encoding and modulation.

(57) A method and apparatus for combining encoding and modulation creates signal sets from available amplitude and phase modulations by indexing ordered subspaces. The subspaces need not be limited to the class of subspaces known as binary subspaces. The resultant signal sets, for a preselected power and bandwidth, are widely separated and unlikely to be confused by the effects of channel noise. Such signals can be in either finite block or convolutional form, depending on the natural format of the desired transmission. Further according to the invention are basic apparatus for encoding and modulating as well as demodulating and decoding a signal in accordance with the invention. Specifically, a method is provided for decoding that incorporates a specific type of decoding/demodulation techniques which develops accurate estimates of the information from the received signal in a computationally efficient manner and which permits high speed operation using soft-decision decoders.

EP 0 282 298 A2

Description

METHOD AND APPARATUS FOR COMBINING ENCODING AND MODULATION

BACKGROUND OF THE INVENTION

This invention relates to digital communications and more specifically to techniques for constructing bandwidth efficient signal sets by combining error correcting encoding with modulation of digital data. More specifically, the invention relates to generalized methods and apparatus for encoding and modulating digital information signals and methods and apparatus for demodulating and decoding signals from an information channel containing information. The invention finds particular application to time-division multiple-access (TDMA) operation in a frequency-division multiple-access environment, such as a satellite transponder channel.

In order to aid in identifying the relevance of the references cited herein, the references cited herein are referred to frequently by the abbreviations following the citations listed hereinbelow.

In electronic data communication systems, random noise or interference can cause the transmitted signal to be contaminated and lead to errors in the received message. In systems where the reliability of received data is very important, error-correcting codes have been used to protect the transmitted message and enable system designers to reduce the effects of noise. Two major schools of thought and associated bodies of theory have emerged for performing this task: algebraic block coding, which relies heavily on the use of modern algebra and typically constructs codes as linear subspaces of a fixed size vector space over a finite field; and convolutional coding, in which the transmission is viewed as being continuous and the design typically relies more on computer search techniques and close analysis of the state diagram of the possible convolutional encoder circuits.

For many years, the coding process was effectively separated from the problem of modulation in conventional systems. Modulation is the creation of, for example, electromagnetic signals in which changes in phase, frequency, or amplitude are used to distinguish different messages.

Referring to Figure 1 representing a prior art system 10, in conventional systems 10 a block or stream of information digits 12 is fed into a digital encoder 14 designed for a specific error-correcting code where redundant check bits are added. The resultant encoded digits 16 are then fed into a modulator 18 where each digit or set of digits is typically mapped to a modulated symbol to be transmitted as information in for example a radio frequency signal 20. The radio frequency signal 20 is applied to a channel 22 wherein noise and interference 24 are added and then received as a signal with errors 26 at a demodulator 28. The demodulator 28 attempts to extract from the signal with errors 26 redundant digits with errors 30 which are fed to an error correcting decoder 32 designed to accommodate the error correcting code. The decoder 28 then uses known redundancy structure of in the encoded digits 16 to eliminate as many errors as possible producing as its output estimated received digits 34. In some systems, the demodulator 28 also provides "soft decision information" or "reliability" information along with the estimate of the received digits which can be used effectively in a variety of error-correcting decoders to improve performance, particularly Viterbi decoders for convolutional codes.

To maintain the separations between different messages guaranteed by the minimum Hamming distance of the error-correcting code, the mapping performed by the demodulator 28 must be chosen with care. (The Hamming distance between two words is the number of digits in which the two words differ. The minimum Hamming distance of a code is the minimum over all pairs of code words in the code of the Hamming distance between the two code words.) For example, in binary systems using phase-shift modulations, the correspondence between the redundant binary sequences and the particular phase of a transmitted signal is often dictated by a Gray code.

The use of error-correcting coding in this manner frequently is an alternative to increasing the power of the transmitted signal to overcome the noise. Conversely, the use of coding permits the power of the transmission to be reduced with no degradation in the reliability of the transmission. The power savings obtained in this way are measured in terms of the allowable reduction in decibels of power-per-bit for the same bit error rate, a quantity referred to as "coding gain." However, since coding requires the addition of redundant digits, for a fixed modulation scheme the use of coding requires that symbols be sent at a faster rate, thereby increasing the frequency bandwidth occupied by the transmission.

As the demand for communication links has increased, there has been growing competition for the available electromagnetic spectrum, and significant expansion of the bandwidth of the signal to reduce the power required has no longer been acceptable in many instances. Thus attention has turned to methods of combining coding and modulation into one coordinated mapping to achieve signals that are efficient in both power and bandwidth utilization. In the past, efforts have followed the two pathways set by error-correcting coding theory, with some building on the concepts of convolutional codes whereas others start from the block code ideas.

In the convolutional school, a major step forward was made by Ungerboeck as described in his paper "Channel Coding with Multilevel/Phase Signals" [ung], in which he pointed out that the Euclidean distance properties of the electromagnetic signal space could be incorporated into the design of a convolutional code

encoder. Figure 2 illustrates the basic structure for comparison with Figure 1. Using the trellis characterization of the encoder, i.e., a trellis encoder 44, information digits 12 are mapped directly to modulated signals 20 so as to add redundancy only when the electromagnetic symbols are likely to be confused. The error-correcting encoder and modulator are combined into a single coder/modulator herein called the trellis encoder 44. The standard Viterbi algorithm for decoding convolutional codes can be readily adapted to a so-called Viterbi trellis decoder 48 to decode the received symbols (signal with errors 26) directly to estimated information digits 34. In adapting the convolutional coding methodology, Ungerboeck chose not to "pursue the block coding aspect because the richer structure and omission of block boundaries together with the availability of Viterbi ML-decoding algorithm [sic] make trellis codes appear to us more attractive for present coding problem [ung,p.58]."

Others have followed Ungerboeck. For example, recently S. G. Wilson has shown a construction for rate 5/6 trellis codes for an 8-state phase-shift keying (8-PSK) system and has found that it achieves an asymptotic gain of 6.2 dB over an uncoded 8-PSK system [wlsn].

Other researchers have pursued the construction of efficient coding/modulation systems from the algebraic block code point of view. Imai and Hirakawa showed how error-correcting codes of increasing strengths can be coupled to increasingly error-sensitive parameters of the signal modulation in both multilevel and mutliphase modulations to give improved performance. Furthermore they explained a staged decoding method in which the most sensitive parameters are estimated first, using the a posteriori probabilities based on the channel statistics and the code structure wherein those estimates are used in later probability calculations to determine estimates for the successively less sensitive parameters [I&h].

Similarly, V. V. Ginzburg has used algebraic techniques to design multilevel multiphase signals for a continuous channel. His methods address the case where the measure of distance in the continuous channel is monotonically related to an additive function of the distances between individual signal components. (Such additivity is commonly assumed in satellite channel models, for example.) He generalized the ideas of Imai and Hirakawa by partitioning the set of elementary modulation signals into carefully chosen subsets that permit the actual channel distance between signals to be associated with the particular subsets in which the signals are found. He then combined a hierarchy of subsets with a matching hierarchy of codes of increasing strength to design signal sets that are guaranteed to have large separations in the signal space. The algorithms he suggested for demodulating and decoding the signals has been given in only abstract mathematical terms: "A rigorous maximum-likelihood demodulation procedure of acceptable complexity may be built only in exceptional cases. A most simple approximate procedure implementing an energy distance D (i.e., one that leads to a correct decision for a noise energy < D/4) may be built as a sequence of integral reception procedures (to be carried out in the order of decreasing levels) for the individual codes that define the signal-system construction, if each of them implements D ...." [gnz].

Most recently, Sayegh [syh] has developed further the methods of Imai and Hirakawa by explicitly defining particular block codes that can be attached to the various levels of a hierarchy which admits to soft- decision decoding procedures, and he has demonstrated some achievable gains using his methods through simulation studies for very short codes. Sayegh's work is notable as well in that he has shown how Imai and Hirakawa's method can be combined with the signal set partitions of Ungerboeck to create combined coding and modulation systems based on several other signal constellations. Sayegh's work represents what is believed to be the most relevant development to the present invention. However, Sayegh does not represent a prior art publication, since publication was less than one year prior to the filing date of the present application.

Other authors [ck&sl] [frny86] have approached the problems of constructing bandwidth efficient signal sets using mathematically defined lattices. Thus, their work is distinguishable as essentially unrelated to the present scheme.

Deficiencies in the Prior Art

To understand the key advances of the present invention, it is important to understand the deficiency of the conventional practice of decomposing the process of creating the radio wave signal into separate processes of error-correcting encoding followed by modulation with, for example in the case of PSK, a Gray code; how the more recent methods ([I&h], [ung] and [gnz]) create a more effective linkage between the code structure and the demodulator and thus improve the quality of the constructed signal set; and how the new methods of the present invention present significant advantages in the effectiveness of the linkage both for creating high quality signal sets and facilitating computation of the estimation of the transmitted information by the receiver.

The preponderance of the literature and theory on the construction of error-correcting codes addresses the problem of creating linear codes (e.g., [v&o, pp. 82-101]), which are well suited for protecting messages to be sent across a symmetric channel. For the symmetric channel, a central index of the quality of a code is its minimum Hamming distance, the minimum number of digits in which any two distinct code words must differ. The mathematical theory of codes such as BCH codes, which are output from an encoder to a modulator, attempts to guarantee that the minimum Hamming distance of the code in the (digital) vector space of the communication channel is as large as possible for a given code rate and block length, since the larger the minimum distance, the more errors that are certain to be correctable by a maximum likelihood decoding algorithm.

However, on the continuous white Gaussian channel, most modulation schemes do not induce a symmetric channel from the perspective of the digital error-correcting coding system. Certain pairs of elementary

modulations are closer in the Euclidean space of the channel than are others, and thus they are more likely to be confused than others. At very high signal-to-noise ratios, the probability of error for a maximum likelihood decoder is directly related to the minimum Euclidean distance separating any pair of distinct transmitted signals. Thus, for an entire coding/modulation system, an important measure is the minimum Euclidean distance.

To illustrate by reference to Figure 3, in the case of phase-shift modulations such as 8-PSK in conventional systems, the use of a Gray code mapping of binary triples to phase states serves the purpose of assuring that the Hamming distance of the error-correcting code is reflected in a reasonable lower bound on the Euclidean distance. In Figure 3 for a specific example, since by definition of a Gray code, the Gray code binary sequences associated with any pair of adjacent modulations differ in exactly one bit, the minimum squared Euclidean distance of a pair of signals in a two-dimensional signal subspace formed by a binary error-correcting code of minimum Hamming distance D and Gray coding must be at least D δ^2, where δ^2 is the minimum squared Euclidean distance between modulation phases.

This scheme is deficient because it creates a potential ambiguity due to the simultaneous occurrence of minimums. Specifically, the minimum Hamming distance separation in binary sequences can be achieved simultaneously with the minimum squared Euclidean distance of the modulation, thus creating a signal set for which a pair of signals is separated by a squared Euclidean distance that is the product of minimums, D δ^2, which is the undesirable ambiguity.

Both Ungerboeck and Ginzburg avoid the possibility of the simultaneous occurrence of minimums by coupling the coding to the modulation via a careful nested partitioning of the available modulations into subsets. In Ungerboeck's language, the mapping of binary sequences to modulations "follows from successive partitioning of a channel-signal set into subsets with increasing minimum distances:

$\Delta_0 < \Delta_1 < \Delta_2 ...$

between the signals of these subsets." Ungerboeck then attaches modulations of particular subsets directly to edges in a trellis encoder for a convolutional code in such a way that the constraints of the trellis prevent the smallest Euclidean distances from occurring on all of the edges of the short paths originating and terminating at the zero state of the trellis. Effectively, this prevents a simultaneous occurrence of minimums in both the redundancy imposed by the trellis and the Euclidean separation of the modulation.

Similarly, Ginzburg, in his hierarchical construction, defines partitions with L levels and associated minimum squared Euclidean distances satisfying:

$\delta_L^2 < ... < \Delta_2^2 < \delta_1^2$

(adapting Ginzburg's language to the present notation). He then associates a different error-correcting code with each level, the code for the 1th level having a minimum distance D_1. (Note the difference between the variable 1 and the numeral 1 herein.) The squared Euclidean distance for the signals thus created must be at least:

$$D \leq \min_{1 \leq l \leq L} (\delta_l^2 D_l).$$

Since the D_1 values are chosen to be larger for the δ_1 that are smaller, this minimum is much greater than the product of the minima. Likewise, the earlier technique of Imai and Hirakawa may be viewed as a special instance of Ginzburg's signal construction method.

In Ginzburg's method, the use of freely created nested partitions creates some limitations. At the 1th level, there are actually a very large number of partitions, a different partition for each of the subsets created in the partition at the (1-1)st partition. Due to the process of successive subdivision, the number of subsets of modulations that must be considered grows exponentially with the level, and the δ_1 associated with the 1th level is actually a minimum over all of the subpartitions at the 1th level.

With regard to prior art decoding circuitry and methods as hereinafter discussed, all of the workers in the field previously cited have recognized the difficulty of decoding these high performance coding modulation schemes. Typically systems in which combined coding and modulation is contemplated are those where soft decision information is available. The most commonly used algorithm is the Viterbi algorithm, which can incorporate soft decision information into the calculation of path metrics. The path metrics are used to determine the best estimate of each of the transmitted bits.

There is nothing in principle that precludes the use of the Viterbi algorithm in cooperation with encoding/modulation systems with encoders according to the invention as described herein. If all such encoders are convolutional, the encoder can take the form of a trellis encoder and the resultant signal can be decoded using the techniques suggested by Ungerboeck for trellis codes. In practice, however, the powerful codes created by the encoders of the invention as described herein can create redundancies which interrelate a large number of transmitted symbols and, when viewed as trellis codes, have an extremely large number of states. The number of states will in most instances make the Viterbi algorithm impractical.

For the encoder/modulators based on block codes, the same problem is encountered. Most techniques for decoding block codes using soft decision information are very complex. For example, the Chase algorithm [ch] can be employed when the expected number of errors is very small, but the computational effort required

grows exponentially when the number of errors to be corrected increases. Similarly, it is possible to contemplate the use of Forney's generalized minimum distance decoding [fmy66], but the complexity of this technique in practice is usually prohibitive.

What is needed therefore is very efficient low-complexity algorithms. The present inventor as described in Tanner [tan81] has described algorithms which as described hereinbelow have potential interest in decoding/demodulation methods and apparatus according to the invention. The Tanner article is therefore incorporated herein by reference and made a part hereof. Tanner's algorithms lead to a wide variety of decoder architectures, many of which are particularly well suited to parallel implementation in large scale integrated circuits. While the algorithms described in the Tanner article (e.g., Algorithm B) do not perform maximum likelihood decoding, they can be used effectively to decode much longer codes than can other methods, either large length block codes or large constraint length convolutional codes, without incurring prohibitive circuit complexity costs. Moreover, the sub-optimal Algorithm B, for example, can incorporate soft decision information into the decoding process for either block or convolutional codes with very little increase in the complexity of the decoding algorithm. In practice, decoding systems based on Tanner algorithms can outperform many Viterbi algorithm-based systems because the sub-optimality of the decoder performance is minor compared to the advantage of using a more powerful coding/modulation design.

In most advanced modulation systems, the demodulator receives a single symbol that can be viewed as a vector in some symbol vector space. It then computes a distance measure for the separation between the received symbol and all of the possible elementary modulations. In so-called "hard decision" demodulators, the distance measure is heavily quantized to two implicit values. The closest elementary modulation is emitted as the best estimate (implicitly at distance 0) and the others are viewed as being at distance 1. In "soft decision" demodulators, the demodulator can put out numbers, typically viewed as quantized representations of real numbers, that indicate the likelihood that the received vector came from each of the possible elementary modulations.

Making optimal use of this "soft decision" information as well as the constraints imposed by the digital code structure is in general very difficult. While it is possible to contemplate the design of a decoder based on Tanner's algorithms that would use all of the probability information provided by the channel for a sophisticated modulation scheme, heretofore nothing has been taught or suggested which could optimize the process in a simple fashion.

Imai and Hirakawa proposed a technique for pure 2M-ary PSK and pure multilevel signaling wherein the most sensitive or least significant bit of the modulation are decoded using calculated probabilities [I&h, p. 373]. The estimates for these least significant bits are then fed into a decoder for the L-th error-correcting code, generally the most powerful code. The final decision bits from this Lth decoder are then used in combination with the channel soft-decision information in providing the probability estimates needed for the (L-1)st decoder. The final decision bits from the (L-1)st stage along with the final decisions from the Lth stage are in turn used in providing the probability estimates needed for the (L-2)nd decoder, and so forth, until the most significant bits are decoded.

While the description is very incomplete, Sayegh's decoding procedures [syh, p. 1044] appear to be those of Imai and Hirakawa adapted to the additional modulations he treats, requiring an optimal decision at each stage of a sub-optimal procedure. This decomposition adds a further source of potential loss. If the decoder at some stage does not provide the correct final decision bits, the estimates used in calculating probabilities for all successive stages will reflect the errors and increase the possibility of decoding error in those later stages. However, for high-speed systems the decomposition has the advantage of creating parallel data paths for the decoding of the bits of different significance. In the representative circuit shown in Figure 2 of Imai and Hirakawa, there are four different decoders for different error-correcting codes all working simultaneously in pipelined fashion. As a result, each of the decoders must handle data at only a fraction of the raw input rate. In high performance systems, very efficient error-correcting codes must create dependencies that interrelate large numbers of digits, requiring either a long block length in the case of algebraic block codes or a long constraint length in the case of convolutional codes. As is well known to those acquainted with the coding arts, decoding such codes using soft decision information is very complex. Typical implementations for such decoders either cannot handle high data rates or require large and complex circuits that consume significant power. What is needed is a more efficient decoding scheme.

SUMMARY OF THE INVENTION

According to the invention, method and apparatus provide combined encoding and modulation which creates signal sets from available amplitude and phase modulations by indexing into ordered subspaces. The ordered subspaces need not be limited to the class of subspaces known as binary subspaces. The resultant signal sets, for a preselected power and bandwidth, are widely separated and unlikely to be confused by the effects of channel noise. Such signals can be in either finite block or convolutional form, depending on the natural format of the desired transmission. Further according to the invention are basic apparatus for encoding and modulating as well as apparatus for demodulating and decoding a signal in accordance with a demodulation/decoding method of the invention. Specifically, a demodulation/decoding method is provided for incorporating a known decoding technique that develops accurate estimates of the information from the received signal in a computationally efficient manner and that permits high speed operation using soft-decision decoders.

The invention will be better understood by reference to the following detailed description in conjunction with the accompanying drawings.

BRIEF DESCRIPTION OF THE DRAWINGS

Figure 1 is a block diagram of a first prior art error correcting code-based communication system.

Figure 2 is a block diagram of a second prior art error correcting code-based communication system.

Figure 3 is a phase diagram (modulation constellation diagram) illustrating a Gray code indexing of phase states for 8 PSK modulation.

Figure 4 is a phase diagram illustrating indexing of phase states for a set partitioning technique.

Figure 5 is a block diagram of a general encoder/modulator structure in accordance with the invention.

Figure 6 is a constellation diagram illustrating indexing of an 8-AMPM signal in accordance with one embodiment of the encoder/modulation technique according to the invention.

Figure 7 is a constellation diagram illustrating indexing of an 8-AMPM signal in accordance with a second embodiment of the encoder/modulation technique according to the invention.

Figure 8 is a phase diagram illustrating indexing of a 6-PSK signal in accordance with a further embodiment of the encoder/modulator according to the invention.

Figure 9 is a flow chart of a generalized decoding/demodulation method implemented in a preferred embodiment of a demodulator/decoder implemented in accordance with the invention.

Figure 10 is a phase diagram illustrating extraction and demodulation of a signal of a prior art decoding and demodulation method.

Figure 11 is a phase diagram illustrating extraction and demodulation of a signal according to one embodiment of the present invention.

Figure 12 is a phase diagram illustrating extraction and demodulation of a signal according to a further embodiment of the present invention.

Figure 13 is a block diagram of an encoder/modulator structure in accordance with a preferred embodiment of the invention.

Figure 14 is a block diagram of a demodulator/decoder in accordance with a preferred embodiment of the invention.

DETAILED DESCRIPTION OF THE INVENTION

According to the present invention, a method is provided for making the minimum Euclidean distance in the Euclidean space of a communication channel as large as possible, or, equivalently and more conveniently, to make the squared Euclidean distance as large as possible, rather than concentrating alone on the Hamming distance of the error-correcting code in the vector space of the digital channel. More specifically, modulations are indexed to form them into ordered subspaces, which include both binary and nonbinary subspaces. The subspaces provide more highly disciplined indexing that could otherwise be exploited without further detailed knowledge of the structure of the particular modulations.

Encoding and Modulation Methods and Circuits

To create the tightest linkage between the structure of the digital error-correcting code and the modulation, the digital indexing of the possible modulations is crucial. Algebraic error-correcting codes are in almost all instances based upon finite fields and can be organized into subspaces. These subspaces are themselves error-correcting codes of potentially different correcting abilities. The well-known BCH codes, for example, can be viewed as nested, with a larger minimum distance code with more numerous roots being a subcode of a code with fewer roots and smaller minimum distance. For instance, the words of the (15,1,15) BCH code are all words in the (15,5,7) code, which are in turn all words in the (15,11,3) code, which are in turn all words in the trivial (15,15,1) code. It has been shown by others that such nested subcode structure could be used to construct recursively long and powerful error-correcting codes based on simple codes that form nested subspaces.

According to the present invention, however, the method of constructing nested subcodes from error-correcting coding theory is adapted to the construction of combined coding and modulation systems. The available modulations are indexed by vectors over a finite field in such a way that the additive modulation distances are well organized by subspaces and affine (shifted) varieties of those subspaces.

To illustrate, with reference to Figure 4, consider the indexing of the 8 PSK modulations induced by set partitioning of the type taught by Ungerboeck. In accordance with the invention, a quanternary phase shift keying set with minimum squared Euclidean distance of 2 is formed by fixing the set of modulations formed by the last of the three bits at either 0 or at 1. Consequently any pair of modulations differing only by changes restricted to the two-dimensional subspace formed by the topmost, the most significant bit, and the center, the center significant bit, have a squared Euclidean distance separation of at least 2. Similarly, if the two rightmost indexing bits, the center and the least significant bit, are fixed at any one of the four possible values, the pair of modulations differing only in the most significant bit have a squared Euclidean distance separation of 4. (In the case of 8 PSK, this indexing of modulations is like that created by Imai and Hirakawa, using different notation.)

Figure 5 illustrates a basic encoder/modulator apparatus 100 according to the invention for exploiting the organization of the modulations described hereinabove. The encoder/modulator apparatus 100 comprises a

first, or lsb, encoder 102, a second, or csb, encoder 104, a third, or msb, encoder 106 and a digit-to-modulation mapping subsystem 108 coupled to receive, respectively, the most significant bit (msb), the center significant bit (csb), and the least significant bit (lsb) from inputs of information digits. The lsb encoder 106 is of the form which generates a strong binary error-correcting code with minimum Hamming distance D_3. The lsb code is the strongest code and therefore has the most impact on decoding. A change in one information bit going into the lsb encoder 106 causes a number of bits at least equal to the value D_3 to be changed going into the mapping circuit (a conventional one-for-one table translation from a pure digit to a radial value signal, which in turn is applied to an analog signal modulator), which in turn leads to a squared Euclidean distance of at least $\delta_3^2 D_3$, no matter what changes occur in the information bits going into Encoders 1 and 2. If no changes occur in the information bits going into lsb encoder 106, then any change in the information bits going into csb encoder 104 must cause a number of bits equal to at least the value D_2 to be changed, which then leads to a squared Euclidean distance of at least $\delta_2^2 D_2$, no matter what changes occur in the information bits going into msb encoder 102. Finally, if no changes occur in the information fed into lsb or csb encoders 106 or 104, any change in the information bits going into msb encoder 102 must cause at least D_1 bits to be changed, which leads to a squared Euclidean distance of at least $\delta_1^2 D_1$. The minimum squared Euclidean distance separation of any pair of encoded modulations is thus at least:

$$D \geq \min_{1 \leq l \leq L} (\delta_l^2 D_l).$$

The method practiced by apparatus according to the invention differs significantly from the methods of the prior art. For example, unlike both Ungerboeck and Ginzburg, the modulations are organized by subspaces, rather than by the mathematically more general but practically less useful hierarchy of subsets. The subspaces create more highly disciplined indexing that can be exploited without further detailed knowledge of the structure of the particular modulations.

Similarly, while Imai and Hirakawa's construction lead to subspaces in the case of several particular types of modulations, the possibility of subspace organization was never recognized and thus no advantage has been suggested of the flexibility afforded by subspace organization.

Still further, while Sayegh's constructions (which include 8-AMPM and 16-QASK as well as PSK modulations) use the modulation indexings suggested by Ungerboeck, the Ungerboeck's labellings would induce one of many possible sequences of nested binary subspaces for the modulations described therein, Sayegh failed to recognize or suggest that any equivalent subspace organization is sufficient or that the subspaces need not be binary in general.

To understand the flexibility of the subspace organization, suppose that any arbitrary modulation can be indexed by subspaces of bits such that changes restricted to s_1 bits forming a Subspace 1 induce a squared Euclidean distance of at least δ_1^2; changes restricted to the s_1 bits forming a Subspace 1 or the s_2 bits forming a Subspace 2 induce a squared Euclidean distance of at least δ_2^2; changes restricted to the s_1 bits forming a Subspace 1, the s_2 bits forming a Subspace 2, or the s_3 bits forming a Subspace 3 induce a squared Euclidean distance of at least δ_3^2; and so forth. We assume $\delta_1^2 > \delta_2^2 > \delta_3^2 > \ldots > \delta_L^2$. Typically such subspace indexings of the modulations can be created by using the groups and subgroups of invariances of the set of elementary modulations. In general the subspaces need not be over extension fields of the binary field; fields of characteristic other than 2 may be appropriate for other signal constellations. Typically, however, extension fields of the binary field are of the greatest practical importance, and we will focus attention primarily on such examples.

To perform the combined encoding and modulation, a first error-correcting encoder 102 for a code with minimum Hamming distance D_1, producing symbols consisting of s_1 bits, is used to produce the successive sets of s_1 bits to be fed into the most significant subspace lines 103 of the digit-to-modulation mapping subsystem 108. Similarly, a second error-correcting encoder 104 for a code with minimum Hamming distance D_2, producing symbols consisting of s_2 bits, is used to produce the successive sets of s_2 bits to be fed into the next most (center) significant subspace lines 105 of the digit-to-modulation mapping subsystem 108. This continues until at the bottom, an error-correcting encoder for a code with minimum Hamming distance D_L, producing symbols consisting of s_L bits, is used to produce the successive sets of s_L bits to be fed into the least significant subspace lines of the digit-to-modulation mapping subsystem 108. In practice, the l^{th} encoder may be either a convolutional encoder or a block encoder for digits in an extension field $GF(2^{s_l})$, or, more simply, s_l copies of a binary error-correcting encoder for a code with minimum Hamming distance D_l (although such copies are usually less efficient for the same level of protection). In cases where the Euclidean distance separations of the l^{th} subspace mirror perfectly the Hamming distance separations of the digital code (as is the case for QPSK modulations), the s_l digits for the l^{th} subspace may be l successive digits from an error-correcting code word.

With this subspace organization, by choosing the error-correcting codes to have minimum Hamming distances that very roughly inversely with the squared Euclidean distance of subspaces, it is possible to create combined coding/modulations with very large channel distance separations at relatively high coding rate and

without bandwidth expansion. Two examples of systems for more subtle modulations will serve to illustrate the signal creation method and distinguish the present invention from the more limited methods of Imai and Hirakawa and Sayegh.

Figure 6 illustrates 8-AMPM modulation as two QPSK modulations 50 and 50′ indexed as two subspaces which are combined with binary unit displacement 52. The resulting indexing 56 is illustrated in the adjacent portion of the figure. The X's correspond to the positions defined by the combined quadrature modulations. The corresponding index value is a three-digit binary number of an eye pattern. Figure 6 thus shows how 8-AMPM modulation can be viewed as the combination of two "large grain" QPSK modulations each defined on respective two dimensional unit radius circles 50 and 50′ shifted by a unit displacement 52 relative to each other. A "small grain" binary subspace determines which of the two QPSK versions circumscribed by the unit radius circles 50 or 50′ is sent. If the QPSK signal is indexed to be broken up into two subspaces, one with squared Euclidean distance of 2, e.g., between 111 and 100 (54) and the other with squared Euclidean distance 4, e.g., between 111 and 110 (56) as shown in the right side of Figure 6, these can be used as the most significant bits of a three bit indexing of 8-AMPM. The least significant bit is the value of the binary unit displacement 52 and is the third bit (the left-most bit of each triplet in the Figure) which determines which of the two displaces QPSK versions 50 or 50′ is sent. The result is a labeling identical to that of Ungerboeck as used by Sayegh.

Figure 7 shows another indexing scheme according to the invention. Figure 7 illustrates 8-AMPM modulation as two Gray code signals constellations 60 and 60′ indexed as two subspaces which are combined with binary unit displacement 62. The resulting indexing 66 is illustrated in the adjacent portion of the figure. The X's correspond to the positions defined by the combined quadrature modulations. The corresponding index value is a three-digit binary number of an eye pattern. Figure 7 thus shows how 8-AMPM modulation can be viewed as large grain QPSK components indexed by a conventional Gray code to form two-dimensional subspaces in which the Hamming distance of a binary error correcting code is mirrored exactly by the squared Euclidean distance of the modulation combined by a binary unit of displacement. As before the choice of binary subspace (selected by binary unit displacement 62) determines which of the two versions 60 or 60′ is transmitted. This indexing 66 is shown in Figure 7, and it is obviously not equivalent to Ungerboeck's indexing (equivalent to 56 of Figure 6). With this indexing 66, the two dimensional subspace 60 or 60′ can be determined by two successive bits from a single binary error correcting code of minimum distance D_1. The third subspace bit can determined by another binary error-correcting code of minimum distance D_2. The squared Euclidean distance separation of any two signals created by the system is at least:

$$D \geqq \min(2D_1, D_2).$$

Because longer error-correcting codes are more efficient than short ones for a given minimum distance, this construction in many instances gives more efficient signal sets than does Sayegh's construction and is therefore a notable advance.

The method of construction according to the invention suggests other alternatives. For example, a 16-QASK can be viewed as the combination of a large grain QPSK with a smaller grain QPSK set of displacements. Each of the QPSK signals can be indexed either as two separate subspaces or as a single two-dimensional subspace.

Further according to the invention, constructions based on nonbinary subspaces may also be implemented. With reference to Figure 8, such a possibility will be illustrated by an unusual modulation, 6-PSK. In Figure 8 6-PSK is shown with an indexing in which a single ternary digit from GF(3) (values 0,1 or 2) determines each 3-PSK modulation 70 or 70′ and a single binary digit selects which of two possible 3-PSK modulations 70 or 70′ is used. This leads to a sequence of subspaces consisting of two subspaces: the first is a GF(3) ternary subspace that selects a signal from a 3-PSK set and the second is a GF(2) binary subspace that selects which of two rotated versions of the 3-PSK signal will be used. Using simple trigonometry and assuming a unit radius circle, it can be determined that the squared Euclidean distance associated with the ternary subspace is 3 whereas that for the binary subspace is 1. Here the squared Euclidean distance separation of any two signals created by the system is at least

$$D \geqq \min(3D_1, D_2).$$

As a concrete instance, the first ternary subspace code could be a ternary Hamming (121,116,3) code while the binary subspace could be (121,93,9) shortened BCH code. In 121 transmitted 6-PSK symbols $116(\log_2 3) + 93 = 183.85 + 93 = 276.85$ bits could be transmitted (as opposed to $2(121) = 242$ bits for QPSK) while achieving an asymptotic gain of $\log_{10}(9/2) = 6.53$ dB. over uncoded QPSK.

The encoding structure described above provides a natural parallelism that is convenient for high speed operation. However, using techniques considered state-of-the-art prior to the present invention, many of the signal sets created in this way could not be effectively decoded without the use of enormously complex decoding circuitry. A second advance of the present invention is to provide practical techniques for decoding the combined coding/modulation signal sets at high data rates and with very practical and cost-effective circuits.

Decoding Circuits and Methods

The present invention makes use of selected Tanner's algorithms as described in the 1981 Tanner article, the contents of which is incorporated herein by reference. In the present invention, the decoding/demodulation attempts to use all of the probability information provided by the channel for a sophisticated modulation

scheme in an extremely simple fashion. For high-speed performance, the preferred embodiment breaks the decoding process into stages corresponding to the decoding of the digits by the same subspaces used in the encoding process.

The work of Imai and Hirakawa also serves as a point of departure for demodulation/decoding. In the preferred embodiment of the demodulator/decoder of the present invention, the methods and circuits of Imai and Hirakawa are improved in three ways. First, decomposition is organized by general subspaces in accordance with the invention and implemented in the encoder/modulator of the present invention, rather than by single bits as suggested for the limited set of modulations treated by Imai and Hirakawa. Subspace organization according to the invention permits the methods of the present invention to be adapted to virtually any modulation scheme.

Second, in the present invention, circuits suggested in Imai and Hirakawa are replaced by "extractors" which, at the i^{th} stage, produce quantized logarithms of the likelihoods:

Pr($S = S_i$ ¦ R received), for each distinct index s_i,

where S_is the elementary modulation indexed by s_i in the i^{th} component: i.e., the <u>most likely</u> elementary modulation is indexed by $s_{i+1}, s_{i+2}, \ldots, s_L$, which are the best estimates for the values <u>of the digits</u> of the more sensitive subspaces produced by the decoder in the previous decoding stages. In contrast, Imai and Hirakawa used the more complex calculation of <u>a posteriori</u> probabilities with "intermediate estimation circuits".

Third, and perhaps most importantly, <u>one or more</u> of the decoder circuits is a decoder implementing one of the graph-based decoding algorithms of the Tanner article [tan81]. In the preferred embodiment, Algorithm B of the Tanner article is used for the first time while supplied with quantized logarithms of probability ratios by the "extractor" of that stage. Hence, the Algorithm B is employed with "soft-decision" information.

Referring to the flow chart of Figure 9, Algorithm B is as follows as adapted for decoding with soft decision information according to the present invention. (The algorithm is described with respect to decoding based on hard decision information on page 541 of the Tanner paper):

1) Establish a formal indexing for the registers requires (Step A): Let R_{ij} be the register associated with bit i, where i = 1,2,...,N, which is accessed by subcode processor j, where j = 1,2,...,S. $R_{ij}(t)$ is the value stored by the register R_{ij} after the t^{th} iteration, and R'_{ij} is a corresponding temporary storage register. Let V_i, where i = 1,2,...,N, be a register storing a value $V_i(0)$ which is of value +1 or -1 if the i^{th} bit was received es a 1 (one) or a 0 (zero), respectively. Let J_i be the index set of the subcode processors accessing bit i, and let I_j be the index set of bits accessed by the subcode processor j.

2) Initialize the registers requires: Load each of the V_i registers with a value provided by the corresponding extractor based on information provided by the channel for the i^{th} bit (Step B). Assign register $R_{ij}(0)$ the value in register $V_i(0)$ for each register R_{ij}, for which j is an element of J_i (Step C).

3) Perform an iterative loop (Step D) comprising a subcode phase (Step D1) and a bit register phase (Step D2): In the subcode phase (Step D1), for each value t from 1 to f, where f is a number of iterations selected as a matter of engineering choice for accuracy and speed (the end of the loop is determined in Step D3), determine temporary value R'_{ij} as follows:

$$R'_{ij} = 1/2\Big[\max_{\substack{\underline{C}\ \varepsilon\ \alpha \\ C_i=+1}} \left(\underline{C}\cdot\underline{R}_j(t-1) - \underline{R}_{ij}(t-1)\right) - \max_{\substack{\underline{C}\ \varepsilon\ \alpha \\ C_i=-1}} \left(\underline{C}\cdot\underline{R}_j(t-1) + \underline{R}_{ij}(t-1)\right)\Big]$$

for each i which is a member of the set I_j, where α is the set of vectors derived from all words in the j^{th} subcode, by replacing each one in a code word with a +1 and each zero in a code word with a -1;

$\underline{C} = (C_{i1}, C_{i2}, \ldots, C_{in})$, each derived from a code word in the subcode by replacing each 1 in the code word by a +1 and each 0 by a -1;

$\underline{R}_j(t-1)$ is the ordered vector of register values

$[\underline{R}_{i_1 j}(t-1), \underline{R}_{i_2 j}(t-1), \ldots, \underline{R}_{i_n j}(t-1)]$

with $i_1, i_2, \ldots, i_n$ a member of the set I_j; and $\underline{C}\bullet\underline{R}_j$ denotes a real vector inner product.

If g/2 is odd, and t = 1, then all R_{ij} values are divided by m, the degree of bit nodes in the graph defining the code (in order to avoid multiply counting bits at the bottom of the tree).

In the bit register phase (Step D2), for each i = 1,2,...,N, the registers for the i^{th} bit are updated as follows:

$$R_{ij}(t) = \sum_{l\varepsilon J_i} R'_{il} - R'_{ij} + V_i(0)$$

3) Make a final decision for the value of the bit (Step E): Using the j^{th} subcode processor, find the vector for

$\underline{C}$ which achieves the maximum value for the real vector inner product of the last iteration in the loop and store the corresponding component of the maximizing subcode word in the corresponding temporary storage register. That is, find:

$\max_{c\ \epsilon\ \omega(e} \bullet \underline{R}_l(f)]$, where f is the floor function of (g-2)/4), and set $R'_{i,j} = C_{l,}$.

This will result in the output value of the l^{th} bit being one if the sum of all R'_{ij}, where j is a member of the set J_i, is greater than zero, and zero if otherwise. In other words, the final value of the i^{th} bit is determined by a majority vote of the best final estimates provided by the subcodes checking the l^{th} bit. (Note the difference between the index value 1 and the index variable 1.) Alternatively, the number of iterations may be preselected according to an engineering preference for precision and speed. A suitable number of iterations for a code of less than length 500 is three.

A specific decoder implementing Algorithm B is described in connecting with Figure 13. Figure 13 is described hereinbelow.

The second distinction above will be more easily understood by considering the process of producing estimates for the values of the digits of the i^{th}subspace starting from the most sensitive L^{th} subspace in the simple case where each subspace is binary. For example, consider the 8-PSK modulation of Figure 4. Referring now to Figure 10, using Imai and Hirakawa's method, if vector $\underline{R}$ is received, to calculate the a posteriori probabilities $Pr(s_3=0 ; \underline{R}$ received) and $Pr(s_3=1 ; \underline{R}$ received). ($; \underline{R}$ received means "given $\underline{R}$ is received") all of the eight distances 81-88 from R to each of the elementary modulations 010,001,111,000,100,011,101,110 shown in Figure 10 must be used.

In the extraction method according to the present invention, (Figure 11) only the two elementary distances, 81, with modulation $s_3=0$, and distance 82, with modulation $s_3=1$, are used, assuming that the channel noise density function is a monotonically decreasing function of distance. If the channel noise density function is a monotonically decreasing function of distance, as is the case for white Gaussian noise, the two indicated distances 81 and 82 will be the shortest distances to the two most likely elementary modulations $s_3=0$ and $s_3=1$, respectively. For relatively high signal-to-noise ratios, the a posteriori probabilities associated with the distances 81 and 82 are dominant in the calculation of the true a posteriori probabilities, and the probability calculation is greatly simplified by the use of these two alone.

Consider an example with reference to Figure 12 and Figure 14 of 8 PSK modulations. (The full structure of a demodulator/decoder is described hereinafter). The decoder (306) for the least significant bit is operative to decide that for example the best estimate for s_3, $\underline{s}_3$, is 0. An extractor (Figure 14 314) according to the present invention at the next stage is operative to calculate the logarithm of the likelihood ratio of only the two most likely elementary modulations. (Imai and Hirakawa transpose most significant bit and least significant bit designations. In Imai and Hirakawa's method, at the next stage the probability computation would have to use the probabilities associated with four distances, rather than only two distances 91 and 92.) The extractor may be specific for the type of noise expected in the channel. For this purpose, the extractor may be a simple read-only memory circuit which produces a defined precomputed output in response to an input based on a limited number of possible inputs. The results are used as the soft decision information for a decoder (Figure 14, 308) which produces a best estimate for $s_2 = \underline{s}_2$. In the case of 8 PSK modulations, the estimates $\underline{s}_2$ and $\underline{s}_3$ are used by the third extractor (320) that provides log-likelihood ratios for the decoder (324) determining the best estimate for s_1, $\underline{s}_1$. Since $\underline{s}_2$ and $\underline{s}_3$ are specified (for example $\underline{s}_2=1$ and $\underline{s}_3=0$), there are only two elementary modulations possible $s_1 = 0$ and $s_1 = 1$, as shown in Figure 12, so the third extractor (320) for the last stage is operative to calculate the same log-likelihood ratio as would Imai and Hirakawa's circuit, except much more efficiently.

It can be appreciated that for complicated multilevel and multiphase modulations, the use of only the most likely elementary modulations represents a substantial simplification of the calculation required by the known prior art.

The third advance is the use of decoders based on specific algorithms, and most specifically Tanner's Algorithm B above. In the preferred embodiment, the most powerful error-correcting code is applied to the digits governing the most sensitive index subspace of the modulation. The graph-based algorithms of Tanner, particularly Algorithm B perform decoding as multiple iterations of replicated simple primitive operations that can be performed in parallel. As a result they can be implemented in large scale integrated circuits which can handle extraordinarily high data rates and decode highly efficient codes with soft decision information using a relatively small number of integrated circuits. This property makes the algorithms an excellent choice for the error-correcting decoders of the present invention, particularly for the most powerful error-correcting code in high performance systems.

Figure 13 illustrates a preferred embodiment of an encoder/modulator 200 according to the present invention for the particular case of 8 PSK modulation on a white Gaussian noise channel at an operating signal-to-noise ratio E_s/N_o of approximately 13.5 dB (output bit error rate of approximately 5×10^{-7}) (compare Figure 5). Figure 14 illustrates a preferred embodiment of a demodulator/decoder 300 according to the present invention for the same type of encoding and modulation. While the encoder/modulator and demodulator/decoder shown here are compatible, there is nothing that precludes the use of an encoder/modulator of a different design with the demodulator/decoder of the invention or the use of a demodulator/decoder of a different kind with an encoder/modulator according to the invention so long as the signal information is in a recognized modulation format.

Referring to Figure 13, the strongest code, for the least significant bit or third incremental subspace of the 8

PSK indexing, is a length 73 perfect difference set code with 45 information bits and minimum distance 10 defined by a 73 by 73 circulant parity check matrix with 10 ones per column and 10 ones per row. The code is shortened by one bit to (72,44,10) by setting one of the information bits to 0 (encoder 206). The code for center significant bit is a [72,63,4] shortened Hamming code formed by using 72 of the weight three columns of the parity check matrix of a length 511 Hamming code as the parity check matrix (encoder 204). The code for the most significant bit is a trivial (72,72,1) code; that is, the information is uncoded (line 202). The encoder thus has the form shown in Figure 5 and is coupled to a digit-to-modulation mapping subsystem 108 wherein 8-PSK modulation is applied to an information channel in the conventional manner. (If the order of bit significance is reversed, the form of the encoder is readily compared with the ordering of subspaces and presentation of Imai and Hirakawa.).

The structure of the demodulator/decoder system 300 according to the invention is shown in Figure 14. The input is a signal represented by the time-related function $r^{(t)}$, which is an analog input signal or an equivalent digital signal representative of the information and noise in the communication channel (for example, a sampled data stream of discrete digital values which have a one-for-one correspondence with the analog time-domain signal at the terminal of the communication channel). The demodulator/decoder subsystem 300 produces as an output the estimated digital subspace vector values $\underline{s}_1^{(t_3)}$, $\underline{s}_2^{(t_3)}$, and $\underline{s}_3^{(t_3)}$, which represent the distinct $\underline{s}_i$ values delayed by "d" time units. Specific time delays are as follows: $t_1 = t - d_1$; $t_2 = t_1 - d_2$; $t_3 = t_2 - d_3 = t - d$; where d_i represents the delay of decoding in accordance with the order of processing for a received vector. Several extraction circuits are employed which are operative to calculate log-likelihood values as described hereinabove or as elsewhere in the art (for example, in Imai and Hirakawa).

The demodulator/decoder subsystem 300 comprises a first intermediate extraction circuit E_1 302 coupled to receive an analog signal or equivalent bit stream $r^{(t)}$ and to apply a first extracted output signal 304 to a first algorithm-specific decoder D_1 306. A second algorithm-specific decoder 308 is also provided. The first and second algorithm-specific decoders 306 and 308 implement the Tanner Algorithm B (Figure 9) and comprise a plurality of fixed registers and working registers (not shown) as needed to execute the algorithm. The algorithm may be implemented in connection with a digital computer processing system of a standard design or of a design especially suited to digital signal processing applications. One application has been described in the Chethik et al. paper incorporated herein by reference.

The output of the first algorithm-specific decoder is a binary digital value $\underline{s}_1^{(t_1)}$ coupled to a first time buffer 310, the output of which is a binary digital value $\underline{s}_1^{(t_2)}$ supplied to a second time buffer 312, the output of which is the desired estimate value of s_1 for the time slot or frame t_3.

The output of the first algorithm-specific decoder 306 is also coupled to a second intermediate extraction circuit 314 to which is also coupled an analog or equivalent input signal $r^{(t_1)}$, which is the input signal r delayed by d_1 time unit via a first input time buffer b_1 316. The second intermediate extraction circuit E_2 314 processes the delayed signal or equivalent bit stream $r^{(t_1)}$ in view of the estimate $\underline{s}_1^{(t_1)}$ and applies a result in nonbinary form to the second algorithm-specific decoder D_2 308. The output of the second algorithm-specific decoder D_2 308 is the estimate $\underline{s}_2^{(t_2)}$, which in turn is coupled to be applied to a third time buffer B_3 318 and to a third intermediate extraction circuit E_3 320. The output of the third time buffer B_3 318 is the desired output estimate $\underline{s}_2^{(t_3)}$.

The output of the second algorithm-specific decoder 308 and the output of the first algorithm-specific decoder 306 are also coupled to the third intermediate extraction circuit 314 to which is also coupled an analog or equivalent (delayed) input signal $r^{(t_2)}$, which is the input signal r delayed by $d_1 + d_2$ time units via the first input time buffer b_1 316 and a second input time buffer b_2 322. The third intermediate extraction circuit E_3 320 processes the delayed signal or equivalent bit stream $r^{(t_2)}$ in view of the estimates $\underline{s}_1^{(t_2)}$ and $\underline{s}_2^{(t_2)}$ and applies the result in nonbinary form to a hard decision decoder H 324 for the most significant bit. The output of the hard decision decoder H 324 is the desired output estimate $\underline{s}_3^{(t_3)}$.

The decoder system 300 operates as follows: The Algorithm B decoder 300 for the least significant bit (D_1 306) loads its fixed registers and working registers (not shown) for each bit with the log-likelihood ratio for that bit provided by the extractor (E_1 302), each log-likelihood represented by a 6-bit quantity in the exemplary implementation. The two-phase correction process of Algorithm B (according to the flow chart of Figure 9) is repeated for example three times before a best estimate for each bit is produced. The best estimates are fed to the extractor for the next stage. The quantized log-likelihood values provided by the extractor for the second stage are fed into the second Algorithm B decoder D_2 308 for the shortened Hamming code. The decoder 308 is structured on a graph for the code in which every bit is checked by exactly three of the nine parities. The resultant two estimates produced $\underline{s}_1$ and $\underline{s}_2$ are fed to the third extractor E_3 320 for the most significant bit stage. At this operating point, no coding is required on the most significant bit inasmuch as the probability of error of binary PSK is lower than the error rate for the estimates produced by the (72,44,10) decoder. Consequently, the extractor 320 output is hard limited by the hard decision decoder 324 to produce a hard decision for the most significant bit. The output of the hard decision decoder H 324 is thus the desired output estimate $\underline{s}_3^{(t_3)}$, which together with the other values $\underline{s}_1$ and $\underline{s}_2$ represent the best estimate of the information transmitted from the remote data source.

Simulation studies of this particular preferred embodiment have been conducted. The system achieved better than 6 dB of coding gain over uncoded 8 PSK with a Gray code mapping of information to modulations at an uncoded bit error rate of 5×10^{-7}. The overall coding rate for the system was $(44+63+72)/(3x72) = 82.8\%$. This compares very favorably with any other system proposed or

implemented. At the finite bit error rate operating point of the present system and with actual implementation losses, a trellis code will achieve less than its asymptotic gain.

The preferred implementation discussed is only illustrative of the many forms that combined coding and modulation systems realizing the advances of the present invention can assume. For example, it should be understood that the application of the invention to non-binary subspaces includes multidimensional subspaces and the term binary subspace as used herein applies to one-dimensional binary subspaces. It is therefore not intended that this invention be limited except as indicated by the appended claims.

Claims

1. In a digital signal communication system for communicating information through an information channel, a method for combined coding and modulation of digital information comprising the steps of:

indexing digital signals representative of elementary modulations by indexing vectors to create a decomposition of indexing vectors of an index vector space into a plurality of ordered subspaces, including binary and nonbinary subspaces;

associating with each said indexing vector a Euclidean distance in modulation space such that any two modulations whose indexing vectors differ only by a distance vector contained in a first subspace and any preceding (higher significant) subspaces of the series of ordered subspaces are separated in said modulation space by at least said Euclidean distance; and

encoding information signals by encoders employing error-correcting codes, each said encoder producing a symbol representative of an indexing vector of the same dimension as a corresponding one of said ordered subspaces for communication of said symbol through said information channel.

2. A method according to claim 1 wherein said error-correcting codes are finite length block codes.

3. A method according to claim 1 or 2 wherein said error-correcting codes are continuous overlapping convolutional codes.

4. A method of encoding the modulated signals according to claim 1, 2 or 3 wherein said encoders encode concurrently to produce redundant indexing vectors, further including the step of providing said redundant indexing vectors to a modulator mapping subsystem concurrently along parallel data paths.

5. In a digital signal communication system for communicating information through an information channel, an apparatus for combined coding and modulation of digital information comprising:

means for indexing digital signals representative of elementary modulations by indexing vectors to create a decomposition of indexing vectors of an index vector space into a plurality of ordered subspaces, including binary and nonbinary subspaces;

means coupled to said indexing means for associating with each said indexing vector a Euclidean distance in modulation space such that any two modulations whose indexing vectors differ only by a distance vector contained in a first subspace and any preceding (higher significant) subspaces of the series of ordered subspaces are separated in said modulation space by at least said Euclidean distance; and

encoding means coupled to said associating means employing error-correcting codes, each said encoding means for producing a symbol representative of an indexing vector of the same dimension as a corresponding one of said ordered subspaces for communication of said symbol through said information channel.

6. An apparatus according to claim 5 wherein said error-correcting codes are finite length block codes.

7. An apparatus according to claim 5 or 6 wherein said error-correcting codes are continuous overlapping convolutional codes.

8. An apparatus according to Claim 5, 6 or 7 wherein said encoding means encode concurrently to produce redundant indexing vectors and further including a modulator mapping subsystem coupled to said encoding means, said modulator mapping subsystem for receiving said redundant indexing vectors concurrently along parallel data paths.

9. An apparatus for encoding and modulating digital information to be transmitted as a signal in an information channel comprising:

a first encoder;

at least a second encoder; and

a digit-to-modulation mapping subsystem coupled to receive encoded information digits from respective outputs of said first encoder and said at least second encoder;

wherein said first encoder generates a strong error-correcting code from a set of possible codes including nonbinary code, with minimum Hamming distance between encoded information digits such that a change in one information digits applied to said first encoder causes a number of digits at least equal to the value of said minimum Hamming distance to be changed for application to said mapping subsystem, with a squared Euclidean distance of at least $\delta^2 D$, where δ is the Euclidean distance and D is the Hamming distance, independent of any changes which occur in the information digits applied to said at least second encoder.

10. A combined decoding/demodulation apparatus for recovering digital information in the form of

indexing vectors in ordered subspaces from an information channel comprising:

means for receiving noise and modulations through the information channel, said receiving means being suited to reception of modulations introduced into said information channel including more than pure phase shift keying and pure multilevel signaling;

means coupled to said receiving means for estimating transmitted information from the received signal, said estimating means including successive stages of circuitry for extracting information pertaining to probability for a value of an indexing vector of a last subspace in a series of ordered subspaces not yet estimated as a function of estimated values of indexing vectors in all previously decoded subspaces; and

means coupled to said estimating means for decoding the value of said modulations as an indexing vector value for an error-correcting code associated with each successive subspace thereby to produce estimates for each indexing vector through a last indexing vector of a last subspace.

11. An apparatus according to claim 10 wherein at least one of said decoding means includes means for computing a plurality of subcodes and means for selecting from said plurality of subcodes a best estimate of a digit value for a received modulation.

12. An apparatus according to claim 11 wherein one of the decoding means comprises a Tanner Algorithm B decoder employing soft decision information.

13. An apparatus according to claim 10, 11 or 12 wherein said extraction means comprises means for producing decoder input information as a function only of a posteriori probabilities for each most likely elementary modulation indexed by each possible indexing vector for its associated subspace in order and also indexed by estimated vectors produced by the decoders for each preceding ordered subspace.

14. In a digital signal communication system for communicating information through an information channel, a method for combined decoding/demodulation to recover digital information in the form of indexing vectors in ordered subspaces from said information channel, said method comprising the steps of:

receiving noise and modulation through said information channel, said modulations introduced into said information channel including more than pure phase shift keying and pure multilevel signaling;

estimating transmitted information from the received signal in successive stages for extracting information pertaining to probability for a value of the indexing vector of e last subspace in a series of said ordered subspaces not yet estimated as a function of estimated values of the indexing vectors in all previously decoded subspaces; and

decoding said modulation value as an indexing vector value for an error-correcting code associated with each successive subspace thereby to produce estimates for each indexing vector through a last indexing vector of a last subspace.

15. A method according to claim 14 further including the step of computing a plurality of subcodes and selecting from said plurality of subcodes a best estimate of a digit value.

16. A method according to claim 15, wherein the decoding step includes applying a Tanner Algorithm B decoding method with soft decision information to at least one said indexing vector estimate.

17. A method according to claim 14, 15 or 16, wherein said extraction step includes producing decoder input information which is a function only of a posteriori probabilities for each most likely elementary modulation indexed by each possible indexing vector for its associated ordered subspace and also indexed by estimated vectors produced by the decoders for each preceding ordered subspace.

18. In a digital signal communication system for communicating information through an information channel, a method for combined decoding/demodulation to recover digital information in the form of indexing vectors in ordered subspaces from said information channel, said method comprising the steps of:

receiving noise and modulations through said information channel, said modulations introduced into said information channel;

estimating transmitted information from the received signal in successive stages and producing decoder input information which is a function only of a posteriori probabilities for each most likely elementary modulation indexed by each possible indexing vector for its associated ordered subspace and also indexed by estimated vectors produced by the decoders for each preceding ordered subspace for extracting information pertaining to probability for a value of the indexing vector of a last subspace in a series of said ordered subspaces not yet estimated as a function of estimated values of the indexing vectors in all previously decoded subspaces; and

decoding said modulation value as an indexing vector value for an error-correcting code associated with each successive subspace wherein the decoding step includes applying a Tanner Algorithm B decoding method with soft decision information to at least one said indexing vector estimate thereby to produce estimates for each indexing vector through a last indexing vector of a last subspace.

FIG.1
(PRIOR ART)

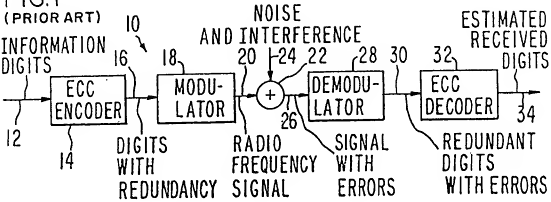


FIG.2
(PRIOR ART)

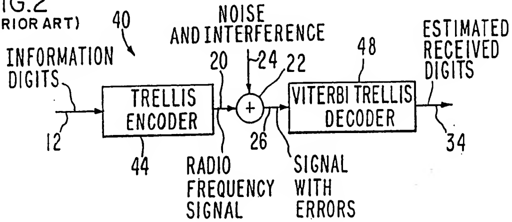


FIG.3
(PRIOR ART)

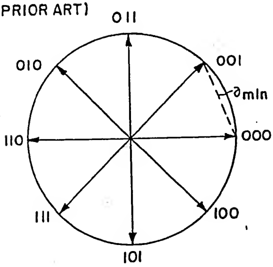
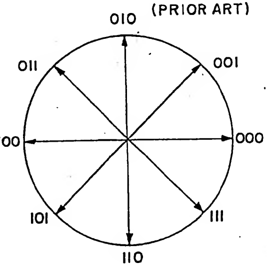


FIG.4
(PRIOR ART)



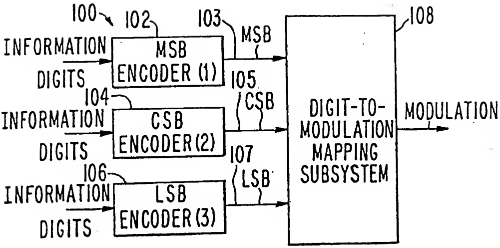


FIG.5

FIG.6

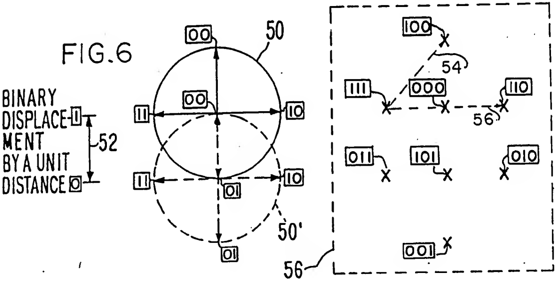


FIG.7

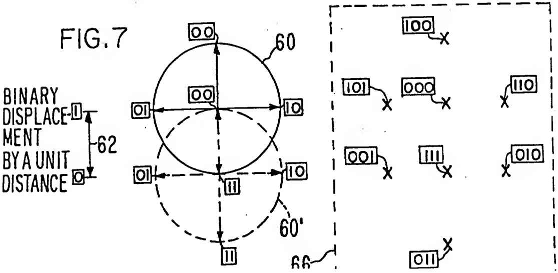


FIG.8

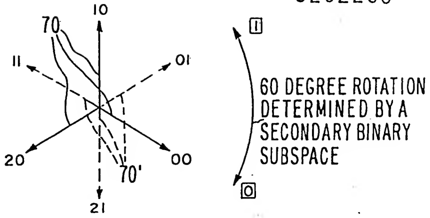

10
70
11
01
60 DEGREE ROTATION DETERMINED BY A SECONDARY BINARY SUBSPACE
1
20
00
70'
0
21

FIG.9

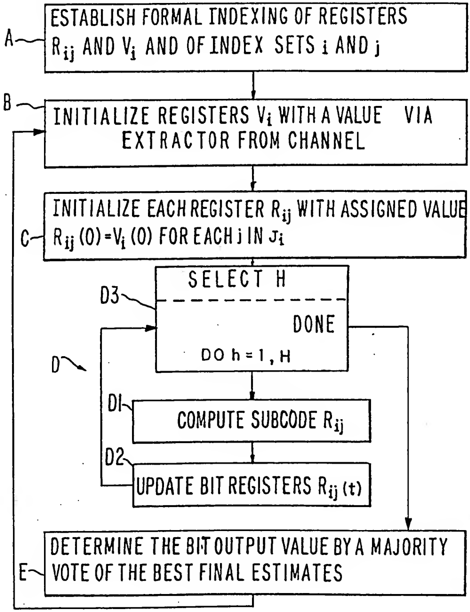

A
ESTABLISH FORMAL INDEXING OF REGISTERS R_{ij} AND V_i AND OF INDEX SETS i AND j
B
INITIALIZE REGISTERS V_i WITH A VALUE VIA EXTRACTOR FROM CHANNEL
INITIALIZE EACH REGISTER R_{ij} WITH ASSIGNED VALUE
C
$R_{ij}(0)=V_i(0)$ FOR EACH j IN J_i
SELECT H
D3
DONE
D
DO h = 1, H
D1
COMPUTE SUBCODE R_{ij}
D2
UPDATE BIT REGISTERS $R_{ij}(t)$
DETERMINE THE BIT OUTPUT VALUE BY A MAJORITY VOTE OF THE BEST FINAL ESTIMATES
E

FIG.10
(PRIOR ART)

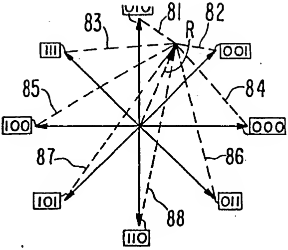


FIG.11

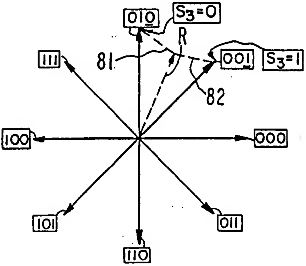
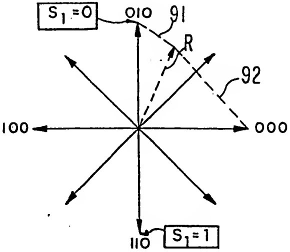


FIG.12



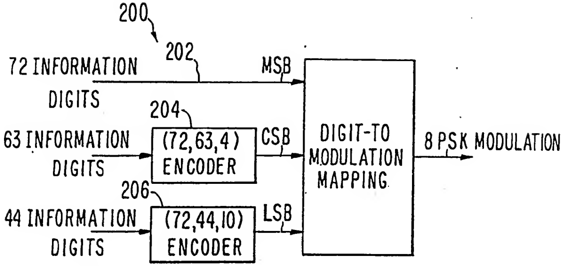
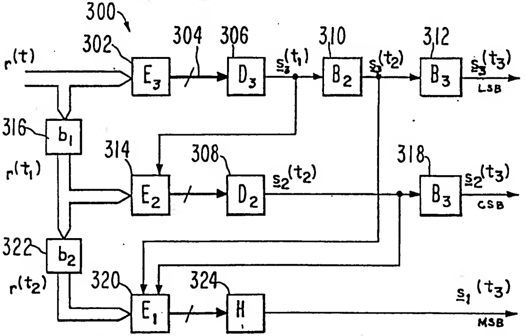


FIG.13

FIG.14



$t_1 = t - d_1, t_2 = t_1 - d_2, t_3 = t_2 - d_3 = t - d$

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(54) Method and apparatus for combining encoding and modulation.

(57) Signal sets are created from available amplitude and phase modulations by indexing ordered subspaces. The subspaces need not be limited to the class of subspaces known as binary subspaces. The resultant signal sets, for a preselected power and bandwidth, are widely separated and unlikely to be confused by the effects of channel noise. Such signals can be in either finite block or convolutional form, depending on the natural format of the desired transmission. Further according to the invention are basic apparatus for encoding and modulating as well as demodulating and decoding a signal in accordance with the invention. Specifically, a method is provided for decoding that incorporates a specific type of decoding/demodulation techniques which develops accurate estimates of the information from the received signal in a computationally efficient manner and which permits high speed operation using soft-decision decoders.

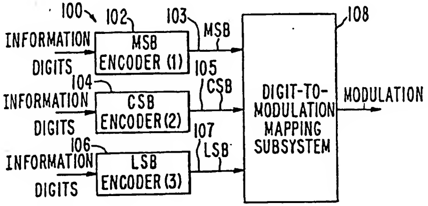

FIG.5

EP 0 282 298 A3

European Patent Office

EUROPEAN SEARCH REPORT

Application Number

EP 88 30 2080
Page 1

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.4)
D,A	IEEE TRANSACTIONS ON COMMUNICATIONS. vol. 34, no. 10, October 1986, NEW YORK US pages 1043 - 1045; SAYEGH S.: "A CLASS OF OPTIMUM BLOCK CODES IN SIGNAL SPACE" * the whole document *	1-3, 5-7, 9-11, 14, 18	H03M13/00 H03M13/12 H04L27/00 H04L27/18
D,A	IEEE TRANSACTIONS ON INFORMATION THEORY. vol. 27, no. 5, September 1981, NEW YORK US pages 533 - 547; M. TANNER: "A RECURSIVE APPROACH TO LOW COMPLEXITY CODES" * the whole document *	1-3, 5-7, 9-18	
D,A	PROC. OF INT. SYMP. OF SPACE TECHNOLOGY OF SCIENCE, TOKYO, 18-23.05.1986 22 May 1986, pages 999 - 1009; MONTE ET AL.: "BANDWIDTH AND POWER EFFICIENT MODULATION AND CODING DEVELOPMENT" * the whole document *	1-3, 5-7	
A	US-A-4077021 (CSAJKA & UNGERBOECK)		**TECHNICAL FIELDS SEARCHED (Int. Cl.4)**
A	EP-A-122805 (CODEX CORP.)		H03M H04L
A	IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE, HOUSTON, 1-4.12.1986 vol. 2, 03 December 1986, pages 1088 - 1094; DIVSALAR ET AL: "MULTIPLE TRELLIS CODED MODULATION (MTCM)"		
D,A	IEEE TRANSACTIONS ON INFORMATION THEORY. vol. 28, no. 1, January 1982, NEW YORK US pages 55 - 67; G. UNGERBOECK: "CHANNEL CODING WITH MULTILEVEL/PHASE SIGNALS"		

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The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	11 JUNE 1991	DEVERGRANNE C.

CATEGORY OF CITED DOCUMENTS

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P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

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EUROPEAN SEARCH REPORT

Application Number

EP 88 30 2080
Page 2

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.4)
D,A	PROC. OF 11th ANN. COMM. SAT. SYSTEMS CONF. SAN DIEGO,CA 16 March 1986, pages 172 - 180; F. CHETHIK ET AL.: "WAVEFORM AND ARCHITECTURE CONCEPTS FOR A HIGH EFFICIENCY TDMA SATCOM SYSTEM" -----		

TECHNICAL FIELDS SEARCHED (Int. Cl.4)

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	11 JUNE 1991	DEVERGRANNE C.

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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A : technological background
O : non-written disclosure
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T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

(19) Europäisches Patentamt
European Patent Office
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(30) Priority: 12.04.89 JP 90623/89

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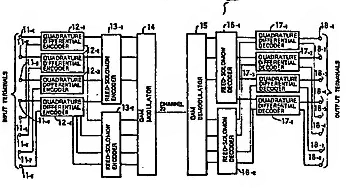
(54) **Quadrature amplitude modulation communication system with transparent error correction.**

(57) In a multi-level QAM communication system, Reed-Solomon encoders and Reed-Solomon decoders are employed for error correction purposes. The phase ambiguity of the received signal is eliminated with differential coding. The multi level QAM communication system (100) utilizing n bits ("n" being an integer) QAM signal having 2^n signal points, comprises: a quadrature differential encoder/decoder unit (12;17) for differentially encoding/decoding n pieces of input digital signal series to produce n pieces of differentially coded signal series; an error correction unit including a Reed-Solomon encoder (13) and a Reed-Solomon decoder (16), provided inside the quadrature differential encoder/decoder unit (12;17) along a signal processing path of the input digital signal series, for error-correcting the n pieces of differentially-coded signal series by utilizing at least one of the digital signal series with employment of a Reed-Solomon code; and, a QAM modulator/demodulator unit (14;15;34;35) for QAM-modulating/demodulating n pieces of error-corrected signal series so as to produce 2^n QAM signals.

FIG.3



EP 0 392 538 A2

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QUADRATURE AMPLITUDE MODULATION COMMUNICATION SYSTEM WITH TRANSPARENT ERROR CORRECTION

Background of the Invention

Field of the Invention

The present invention generally relates to a multi-level QAM (quadrature amplitude modulation) system for transferring a digital signal by utilizing the multi-level quadrature amplitude modulation. More specifically, the present invention is directed to a QAM communication system capable of increasing signal transmission reliabilities by employing a transparent error correcting method.

Description of the Related Art

In a multi-level quadrature amplitude modulation (QAM) communication system in which multi-bit data such as 4 bits data and 8 bits data are transferred with reference to one signal point on a phase plane coordinate including 2^n ("n" being the data bit number) signal points and original data are reproduced based upon the relationship between the amplitude and phase, an utilization efficiency in a frequency becomes high so that this QAM communication system has been widely utilized in digital microwave communications and digital mobile communications.

As previously stated. the signal transmission of the multi-level QAM communication system is carried out with employment of the QAM signals produced by synthesizing two orthogonal I-channel and Q-channel corresponding to each m-level amplitude-modulated signal. Each of these multi-level QAM signals owns m^2 ($= 2^n$) pieces of signal points. For instance, if "m" is selected to be 16 ($n = 8$), this multi-level QAM signal is equal to 256 pieces of QAM signals having 256 signal points.

In a QAM type receiving system employing synchronous demodulation, a carrier wave is first reproduced from this multi-level QAM signal, and then demodulated by utilizing 2 orthogonal-reproduced carrier waves having different phases with each other at 90° (degrees), and thereafter "n" pieces of digital signal series are obtained in total by way of the multi-level identification. In general, there is a drawback in this QAM receiving system that the phases of the reproduced carrier waves derived from the carrier wave reproducing circuit have a so-called "phase ambiguity", i.e., the phase becomes any of 0°, 90°, 180°, and 270°. Generally speaking, since transmission signal series cannot be correctly reproduced if the phase ambiguity exists, it is required to employ any means for eliminating the adverse influences caused by this phase ambiguity. To this end, there are some solutions to resolve such a phase ambiguity. That is, for instance, a known signal series is periodically transmitted, whereas the phases of the reproduced carrier waves are discriminated based upon the relationship between this known signal series and the signal which has been demodulated and judged by the reproduced carrier waves having the phase ambiguity at the signal reception side. Otherwise, a transmission information signal is differential-encoded so as to be transmitted, namely which does not directly correspond to the transmission phase, but corresponds to a relative phase difference of a continuous transmitting symbol. At a signalreception end, when this differential-encoded signal is differential-encoded after being demodulated by the reproduced carrier waves, the phase ambiguity in the reproduced carrier waves can be resolved. In general, since 1 bit error is expanded to a continuous 2-bit error, the differential encoding/decoding method has such an advantage that a circuit arrangement thereof is simple, although the bit error rate in the received signal series is increased as compared with that of the first-mentioned solution method for judging the absolute phase. Moreover, to suppress an increase of a bit error rate caused by a differential coding method, there is another method that a signal point mapping of a QAM signal is a quadrant symmetry mapping. In accordance with the last-mentioned method, since the judgement on the upper 2 bits of the input digital signals which is determined by the orthogonal axes (i.e., I-axis and Q-axis) on the phase plane is adversely influenced by the phase ambiguity, the differential coding operation is required. However, the judgement on other bits thereof which is determined by the respective amplitude levels of the I-axis and Q-axis, is not adversely influenced by the phase ambiguity, so that no differential coding operation is required.

Although the QAM modulation method has an advantage of the higher frequency utilization, there is a drawback that when the number of the bits transmitted with 1 symbol, namely the value of "n" is increased, the bit error rate deteriorated even when the transmission power per 1 bit is selected to be equal. Under

such a circumstance, it is required to improve the bit error rate in the multi-level QAM communication system to employ an error correcting method. On the other hand, a QAM modulation system is originally employed so as to increase the frequency utilization efficiency, and accordingly, there is a severe restriction in an available frequency band in systems with employment of the QAM modulation method, such as a digital microwave radio communication system. As a consequence, it is expected to utilize a higher coding rate having a less redundant bit to be added to the input digital signal in the error correcting method.

Furthermore, various limitations are provided so as to apply the error correcting method to the QAM communication system with employment of the above-described differential coding system. First, when the error correcting encoder/decoder are provided outside the differential encoding/decoding processors, since the 1 bit error occurring on the signal transmission channel is expanded to the 2-bit error due to the differential decoding process, the loads required for the error correcting encoder/decoder become large. In other words, the error correction codes having the greater correction capability are required so as to achieve the same reliability as that of the other case where the error correcting encoder/decoder are provided inside the differential encoding/decoding processors. As a result, since the redundant bit number to be added to the input digital signal is increased, there are problems that the resultant utilization efficiency of frequency is lowered and the circuit arrangement of the error correcting decoder becomes large.

It should be understood that the expression "outside" and "inside" described above are defined as follows. That is, for instance, the error correcting encoder/decoder are positioned outside the differential encoding/decoding circuits in a circuit arrangement provided along a flow of an input digital signal (i.e., along a signal processing sequence).

Conversely, in case that the error correcting encoder/decoder are provided inside the differential encoding/decoding circuits along the signal processing path, the adverse influence caused by the phase ambiguity in the reproduced carrier waves is not yet resolved at the input unit of the error correcting encoder. As a result, in such a case, it is required to employ such an error correcting code, namely a transparent error correcting code that even if the input signal is adversely influenced by the phase ambiguity in the reproduced carrier waves, e.g., bit inversion of the input signal, the error correction can be correctly performed with respect to the bit-inverted input signals.

As an error correcting code, there are a binary error correcting code and a nonbinary error correcting code. When a transparent binary error correcting encoder is employed inside differential encoding/decoding circuits, the transparency can be established by employing error correcting encoders/decoders in "n" pieces of signal series. However, this system has a drawback that when the multiple number of the QAM system is increased, a total number of the required error correcting encoders/decoders is also increased. In addition, there is a drawback in the binary error correcting code such that it is very difficult to produce a code whose coding rate is extremely high. When the decoding delay time of the error correcting code is, for instance, 63 symbols, even the resultant coding rate of the binary BCH (Bose-Chaudhuri-Hocquenghem) codes (63, 57) is 90.5%, by which a single error can be corrected, and thus the frequency band is expanded by approximately 10%. On the other hand, when the nonbinary error correcting code is employed, many difficulties may occur in realizing the above-described transparent conditions. Although there has been proposed that the signal point mapping of the QAM signal is the natural binary mapping and the Lee error correcting code is employed, since only such a case that errors occur in the signal points near the transmission signal points can be corrected based upon the Lee error correcting code, the error correcting effect cannot be expected in the communication channel or path which are subjected to a fading phenomenon. In addition, the coding rate of the Lee error correcting code is not always good as other nonbinarycodes.

As previously described, in the conventional QAM communication system employing the binary error correcting code, there are such problems that since the coding rate cannot be high, the efficiency in the frequency utilization is lowered and also the total number of the required error correcting encoders/decoders to perform the differential encoding operation is necessarily increased. Furthermore, in accordance with the conventional QAM communication system employing the Lee error correcting code, there are such a drawbacks that the error correction can be limitedly executed only to the signal points having a small distance on the signal point mapping.

The above-described problems of the conventional multi-level QAM communication system will now be described in detail.

That is, while the original data is reproduced from the received signal in the conventional multi-level QAM communication system, since the capture phase of the reproduced carrier wave has such phase ambiguity of 0, $\pi/2$, π or $3\pi/2$ radians, the two digital signal series to determine the quadrant of the phase plane are generally differential-encoded/decoded by employing the quadrant differential encoder/decoder.

On the other hand, there exist a natural binary mapping method, a Gray code mapping method and a

quadrant symmetry mapping method as a signal point mapping method for mapping 2^n pieces of signal points from the n bits digital signals.

As typical examples, Fig. 1 represents a signal point mapping for a 16-QAM communication system with employment of the Gray mapping method, whereas Fig. 2 represents another signal point mapping for a 16-QAM communication system with employment of the quadrant symmetry mapping. Further, Fig. 14 indicates a signal point mapping with employment of the natural binary mapping.As apparent from Fig. 1, the respective signal points are symmetrically positioned with respect to the respective I and Q coordinate axes in the Gray coded mapping. To the contrary, the signal points positioned in the respective quadrants are arranged in the quadrant symmetry mapping in such a manner that these signal points are rotated with respect to those of the adjoining quadrants.

In these mapping methods shown in Figs. 1, 2 and 14, the influences caused by the phase shifts of $\pi/2$, π , and $3\pi/2$, which are given to the received signal series, are expressed in Figs. 15A to 15C:

In general, it is known that the transmission capacity and frequency utilization efficiency in such a multi-level QAM communication system can be increased by increasing the signal points. However, the more bit numbers are increased, the more bit error rate is increased due to the imperfectness of the appliances. It is desired that the error correction encoding/decoding operations must be performed with slightly lowering the frequency utilization efficiency so as to improve the QAM communication quality.

Thus, as previously stated, in case that the error correction encoder/decoder are provided outside the differential encoding/decoding circuits along the signal processing path, since the continuous bit errors are produced by the differential encoding operation, the error correcting capability of the error correction code must be emphasized or the interleaver must be employed.

However, when the error correcting capability of the error correction code is increased, the frequency utilization efficiency is deteriorated. When the interleaver is newly employed, not only the circuit scale of the entire system becomes large, but also the decoding delay time is increased. As a consequence, it is generally accepted to arrange such error correction encoder/decoder inside the differential encoder/decoder.

It should be noted that when the error correcting encoder/decoder are arranged inside the differential encoder/decoder, the error correction must be correctly performed even when the signal series are varied as represented in Fig. 14 due to the ambiguity of the capture phase in the reproduced carrier wave, and simultaneously, the phase ambiguity must be preserved even when the error correction encoding/decoding operations are carried out. It should be also noted that the error correction code which can satisfy such a condition is called as a transparent code with respect to a phase rotation in an input signal.

As conventional circuit arrangements for the transparent codes with respect to the phase rotations in the input signals, there have been proposed: Japanese KOKAI (Disclosure) patent application No. 63-210252, and "6GHZ 140MBPS DIGITAL RADIO REPEATER WITH 256QAM MODULATION" by Y. Yoshida et al. Proceedings of International Conference on Communications 1986, No. 46-7, pages 1482 to 1486.

In the multi-level QAM communication system as disclosed in the above-described Japanese KOKAI patent application No. 63-219252, there are various drawbacks. That is, since the error correction encoding/decoding operations are independently performed with respect to each of "n" pieces of digital signal series which constitute the in phase channel and also the channel orthogonalized the in-phase channel, "n" pieces of encoders and also of decoders are required. As a result, the scale of the entire apparatus becomes large.

On the other hand, in the multi-level QAM communication system as described in the above publication, i.e., ICC '86,-46-7, there is employed such an encoding/decoding method with employment of the Lee error correction code, for the respective signal series combinations between n/2 series combinations to constitute the in-phase channel and n/2 series combinations to constitute the orthogonalized channel. However, this conventional communication system is limited to such a natural binarymapping method for mapping the n bits data to the signal points. Furthermore, there are many other limitations for the constituting methods of the error correction codes.

Summary of the Invention

The present invention has been made in an attempt to solve the conventional problems, and therefore has a primary object to provide a QAM (quadrature amplitude modulation) communication system capable of realizing a higher coding rate and higher reliability.

Moreover, the present invention has a secondary object to provide a multi-level QAM communication system in which both the error control code and mapping methods are freely selected, a total quantity of

encoders/decoders is smaller than a bit number of input digital data, and a transparent error correction coding for a phase rotation can be realized.

In addition, a third object of the present invention Is to provide a multi-level QAM communication system in which clock frequencies of error correction encoder/decoder can be lowered with respect to a modulation frequency of a quadrature amplitude modulator.

A quadrature amplitude modulation system, according to the present invention, comprises:
differential encoder/decoder means (12;17) for differentially encoding/decoding n pieces of input digital signal series to resolve phase ambiguity contained in the differentially encoded input signal series;
error correction means including a Reed-Solomon encoder (13;83) and a Reed-Solomon decoder (16;87), provided inside said differential encoder/decoder means (12;17) along a signal processing path of said input digital signal series, for error-control-encoding/decoding said n pieces of differentially-coded signal series by utilizing at least one of said digital signal series to correct errors with employment of Reed-Solomon codes; and,
QAM modulator/demodulator means (14;15;34;36;80;82) for QAM-modulating/demodulating n pieces of error-control-coded signal series so as to produce 2^n QAM signals.

Brief Description of the Drawings

For a better understanding of the present invention, reference is made to the following detailed descriptions in conjunction with the drawings, in which:

Figs. 1 and 2 schematically illustrate known signal point mappings;

Fig. 3 is a schematic block diagram of a QAM (quadrature amplitude modulation) communication system 100 employing a first basic idea, according to a first preferred embodiment of the present invention;

Fig. 4 is a schematic block diagram of another QAM communication system 200 employing the first basic idea, according to a second preferred embodiment of the present invention;

Fig. 5 is a schematic block diagram of an internal circuit of the Reed-Solomon encoder 16 employed in the second QAM system 200;

Fig. 6 is a schematic block diagram of an internal circuit of the Reed-Solomon decoder 16 employed in the second QAM system 200;

Fig. 7 is a schematic block diagram of an internal circuit of the syndrome generator 50 employed in the second QAM system 200;

Fig. 8 is a schematic block diagram of a 256-QAM communication system 300 employing a second basic idea, according to a third preferred embodiment of the present invention:

Fig. 9 is a schematic block diagram of another 256-QAM communication system 400 employing the second basic idea, according to a fourth preferred embodiment of the present Invention;

Fig. 10 is a schematic block diagram of a still further 156-QAM communication system 500 arranged by utilizing the second basic idea, according to a fifth preferred embodiment of the present invention;

Fig. 11 is a schematic block diagram of a 64-QAM communication system 600 constructed by using the third basic idea, according to a sixth preferred embodiment of the present invention;

Fig. 12 is a schematic block diagram of another 64-QAM communication system 700 employing the first basic idea, according to a seventh preferred embodiment of the present invention;

Fig. 13 is a schematic block diagram of another 256-QAM communication system 800 employing unique word adder/detector and no quadracture differential encoder/decoder, according to an eighth preferred embodiment of the present invention;

Fig. 14 schematically illustrates a natural binary mapping; and,

Fig. 15A to 15C are tables for explaining phase reference error effects occurred in the three typical mapping methods.

Detailed Description of the Preferred Embodiments

BASIC IDEAS

Before describing various preferred embodiments, two basic ideas of the present invention will now be summarized.

A multi-level QAM (quadrature amplitude modulation) communication system according to the first basic idea of the present invention, is featured by employing error correcting means for performing both encoding and decoding operations of the Reed-Solomon code under the condition that all or a portion of "n" pieces of input signal series for determining a signal point mapping is used as a symbol. Also, in case that the differential coding operation is performed by employing a natural code mapping, such an error correcting means is employed to independently perform both encoding and decoding operations for the Reed-Solomon code with respect to two "I" and "Q" channels orthogonalized with each other. Although there are many generator polynomials for constracting the Reed-Solomon code, such a Reed-Solomon code that codeword polynomials are not divisible by x-1 (namely, the generator polynomial is not divisible by "x - 1") is utilized so as to establish transparency.

In the above-described first QAM communication system, when (u,k) linear block codes are employed, a frequency band width expanding rate for influencing the QAM communication system is determined by the coding rate of the block codes. To correct a t-symbol, all of the linear block codes must satisfy a limit formula of $t \leq (u-k)/2$. In other words, a redundant symbol number (u-k) cannot be reduced by two times of the correction capability. A Reed-Solomon code can satisfy this limit formula, so that the frequency expanding rate can be suppressed to a minimum value in such a QAM communication system employing a Reed-Solomon code.

Due to the phase ambiguity ol the reproduced carrier waves, the respective "I" and "Q" channels cause signal changes different from each other. However, this adverse inlluences can be eliminated by independently performing both the encoding/decoding operations of the Reed-Solomon code with respect to two "I" and "Q" channels orthogonalized with each other.

Furthermore, even when the signals are inverted in the QAM communication circuit due to the phase ambiguity of the reproduced carrier waves, the transparency can be established by utilizing such a Reed-Solomon code in which a generator polynomial is not diylsible by x-1.That is, in case that the signal point mapping corresponds to the natural code mapping, a necessary/satisfactory condition such that a Reed-Solomon code is equal to a transparent code, is as follows: Any codeword polynomial of the code is not divisible by x-1. Another multi-level QAM communication system according to a second basic idea, owns the following features:

In a multi-level QAM communication system in which a bit number ol transmitted/received data is equal to "n" and there are provided 2^n pieces of signal points, an error correction coding operation is separately carried out with respect to each of signal series used for determining a quadrant of a phase plane, and also to other signal series among "n" pieces of signal series for determining a signal.

As represented in Fig. 15(C), in a multi-level differential QAM communication system in which a signal point mapping is determined based upon a quadrant symmetry mapping, a bit inversion and a signal series substitution may occur with respect the signal series (I_1, Q_1) for determining quadrants of a phase plane when the capture phases of the reproduced carrier waves are shifted by $\pi/2$, π, or $3\pi/2$. However, such a phase ambiguity of the reproduced carrier wave gives no influence to other signal series (I_2, ---, $I_{n/2}$; Q_2,---,$Q_{n/2}$). Therefore, if an error correction coding operation has been performed for the signal series (I_1, Q_1) separately, whereby data bits of which have been inverted can be decoded, even when an arbitrary error correction coding operation is carried out for other signal series (I_2,---, $I_{n/2}$; Q_2,---, $Q_{n/2}$), the transparency of this arbitrary error correction coding operation can be compensated.

Furthermore in another multi-level difference QAM communication system in which a signal point mapping is determined by the Gray coding operation, as represented in Fig. 15(D), a bit inversion and a signal series substitution may occur similar to the previous QAM communication system with respect to the signal series (I_1, Q_1) used for determining quadrants of a phase plane if the phases of the reproduced carrier waves are shifted. To the contrary, in other signal series, only a signal series substitution may occur between the signal series of the I-channel (I_2,---, $I_{n/2}$) and the signal series of the Q-channel (Q_2,---, $Q_{n/2}$). Therefore, such an error correction coding operation by which data whose bit has been inverted can be decoded is independently performed as to the signal series (I_1, Q_1), and an arbitrary error correction coding operation is carried out for each of combinations between other I-channel signal series and the Q-channel signal series. Thus, this coding method can establish the transparency with respect to the phase rotation.

A third basic idea of the present invention is as follows. That is, "h" pieces of input signal series among "n" pieces of input signal series having 2^n QAM signal points are encoded by the Reed-Solomon code on Galois field GF(2^l), where "n" and "h" are integers, "n" is larger than or equal to "h", and "l" is larger than or equal to "h". When a relationship "$l = A \times h$" is satisfied, the clock frequency of the error correcting encoder/decoder can be reduced by 1/A of the QAM modulation velocity. Also, in case of 2 h n (namely, a plurality of signal series are encoded with one Reed-Solomon code, a total number of the combinations constructed of the error correcting encoder/decoder can be selected to be smaller than the total number (n).

of the input signal series.

ARRANGEMENT OF FIRST QAM COMMUNICATION SYSTEM

In Fig. 3, there is shown an arrangement of a QAM (quadrature amplitude modulation) communication system 100 according to a first preferred embodiment of the present invention. This first QAM communication system 100 utilizes the above-described first basic idea therein, and the natural binary mapping method as represented in Fig. 14 (i.e., 16-QAM signal point mapping).

For a better understanding of this natural binary mapping method, the phase reference error effects caused by this mapping method are shown in Fig. 15A in comparison with those of other mapping methods shown in Figs. 15B and 15C. However, such phase errors can be corrected by the QAM communication system 100 (will be discussed in detail).

The modulator transmits each n-bit information symbol by modulating a pair of orthogonal carriers, called I and Q. Each carrier takes one of the $2^{2/n}$, amplitude levels, each representing a set of n/2 information bits, $\{I_1, I_2, ..., I_{n/2}\}$ or $\{Q_1, Q_2, ..., Q_{n/2}\}$. The demodulator regenerates the I-Q carrier with 0°, 90°, 180°, or 270° phase ambiguity.

Fig. 15 shows the phase reference error influence on the received data for typical three signal mapping methods: natural binary, Gray, and quadrant symmetry. In the table, x_i's and y_i's are transmitted bits in the sets $\{I_i\}$ and $\{Q_i\}$, respectively, where x_i, y_i — {0, 1} and i = 1, 2, ..., m. For a differentially encoded multilevel QAM system, an error control scheme must be designed, taking into consideration such phase ambiguity influences.

In the first QAM communication system 100 shown in Fig. 3, 8 digital signal series input into input terminals 11-1 to 11-8 are processed by a quadrature differential coding operation in four quadrature differential encoders 12-1 to 12-4. This differential coding operation is carried out for each combination of two signal series, for instance, the respective combinations of two input signal series 11-1 and 11-5; series 11-2 and 11-8; series 11-3 and 11-7; series 11-4 and 11-8. Two output signals derived from the respective quadrature differential encoders 12-1 through 12-4 are supplied to two Reed-Solomon encoders 13-1 and 13-2 respectively. It should be noted that the generator polynomial of the first Reed-Solomon encoder 13-1 is identical to that of the second Reed-Solomon encoder 13-2. A code employed in the respective encoders 13-1 and 13- 2 is a code of GF (2^4), and a generator polynomial G(X) thereof is not divisible by X-1. In the first and second Reed-Solomon encoders 13-1 and 13-2, the input 4-bit signals are encoded as 1 symbol. Assuming now that the generator polynomial corresponds to $G(X)=(X-\alpha)(X-\alpha^2)$, where is a primitive element of GF(2^4), redundant 2 symbols are added to each of the input 13 symbols. Both the first and second Reed-Solomon encoders 13-1 and 13-2 output digital signals for constituting the I-axis(channel) and Q-axis to a QAM modulator 14. In the first preferred embodiment, the digital signal input into the I-axis is "x", whereas the digital signal input into the Q-axis is "y". The QAM modulator 14 modulates a signal with natural binary mapping, and transmits the modulated digital signal(QAM signals) to the signal transmission channel 20. A QAM demodulator 15 reproduces a carrier wave from the quadrature amplitude modulated (QAM) signal received via the signal transmission channel 20 form the QAM modulator 14 so as to demodulate the input QAM signal by this carrier wave. As previously described, the digital signals input into both the I-axis and Q-axis of the QAM modulator 14 are not always coincident with the output signals of the I-axis and Q-axis of the QAM demodulator 15 due to the ambiguity of phase. When the phase shifts in the reproduced carrier waves are 0°(degree), 90°, 180°, and 270° respectively, the output signals from the QAM modulator 15 are: (I,Q) = (x,y) , (x, $\overline{y}$) , ($\overline{x}$, $\overline{y}$) and ($\overline{x}$,y) respectively, as represented in Fig. 15(a). The output signal of the I-axis in the QAM demodulator 15 is supplied to a first Reed-Solomon decoder 16-1, whereas the output signal of the Q-axis in the QAM demodulator 15 is furnished to a second Reed-Solomon decoder 16-2. That is, the output signal derived from the first or second Reed-Solomon encoders 13-1 and 13-2 is directly input into the first Reed-Solomon decoder 16-1. Otherwise, the above-described output signal is once bit-inverted and the resultant bit-inverted signal is supplied to this first Reed-Solomon decoder 16-1. The second Reed-Solomon decoder 16-2 receives the output signal derived from either the first Reed-Solomon encoder 13-2 in the similar condition as described above. As the first Reed-Solomon encoder 13-1 has the same generator polynomial as that of the second Reed-Solomon encoder 13-2, the following condition is required so as to establish that the error correction encoder/decoder, i.e., the Reed-Solomon encoder/decoder are transparent. That is to say, even when all of the bits of the transmitted code words are inverted, the transparency can be established if the complement of a valid codeword is a valid codeword. As will be discussed later, if the generator polynomial of the code word generated in both the Reed-Solomon

encoders 13-1 and 13-2 is not divisible by x-1, the desirable code words can be obtained even when all of bits of code words are inverted. As a consequence, a 1-symbol error occurring on the signal transmission channel 20 can be corrected in both the Reed-Solomon decoders 16-1 and 16-2. Thus, the outputs derived from the Reed-Solomon decoders 16-1 and 16-2 are furnished to four quadrature differential decoders 17-1 to 17-4 under such a condition that the corresponding signal series are combined. In the quadrature differential decoders 17-1 to 17-4, the quadrature differential decoding operation is carried out for these input signals so as to reproduce desirable signals which will be then output from output terminals 10-1 to 10-8.

In accordance with the first preferred embodiment, there is a particular advantage that the 1 symbol error can be realized at a coding rate of 87%. In comparison with the conventional single-error-correction binary BCH(Bose-Chaudhuri-Hocquenghem) code under the same delay time condition, it becomes (15, 11) code so that the coding rate becomes merely 73%. To the contrary, as previously described, there is another particular advantage that the frequency can be utilized in a considerably higher efficiency.

DESIRED CODE WORD

Even when all of bits of an input signal are inverted, a valid code word can be obtained if a generator polynomial is not divisible by x-1, which will be certified as follows. Considering a Reed-Solomon code of GF (2^s), it is assumed that that $m = 2^s-1$ and a primitive element of GF(2^5) is α. Also, it is asumed that among symbols α^0 to α^m constructed of s bits, $\alpha^b = (1,1,\text{---},1)$. The Galois Field GF($2^5$) is an extension field of GF(2), which results in $2 \cdot X^i = 0$. As a consequence, it can be expressed by $(X^m - 1) = (X - 1) \cdot (X^{m-1} + X^{m-2} \text{---} + X + 1)$. Then, X^m-1 contains as a factor the generator polynomial G(X). As a consequence, if G-(X) does not contain $X = 1$ as a root, since the generator polynomial $(X^{m-1} + X^{m-2} \text{---} + X + 1)$ all coefficients of which are $\alpha^{\emptyset} = (0,\text{---},0,1)$ includes G(X) as the factor, the codeword corresponding to all one polynomial is a valid codeword. The all α^b polynomial $(\alpha^b X^{m-1} + \alpha^b X^{m-2} + \text{---} + \alpha^b X + \alpha^b)$, all bits of which are 1, namely all coefficients are "α^b", may be rewritten as $\alpha^b \cdot (X^{m-1} + X^{m-2} \text{---} + X + 1)$, so that it contains G(X) as a factor and thus the codeword all bits of which are 1 is a validcode word. Accordingly, since the received signal, all bits of which have been inverted, is equal to a signal obtained by adding the code word corresponding $(\alpha^b X^{m-1} + \alpha^b X^{m-2} \text{---} + \alpha^b X + \alpha^b)$ to the transmitted code word S(X), this signal becomes a code word. It should be noted that the generator polynomial of the Reed-Solomon code is given by

$$G(X) = \prod_{i=r}^{d-2+r} (X - \alpha^i) = (X - \alpha^r)(X - \alpha^{r+1})\text{---}(X - \alpha^{d-2+r})$$

,and this generator polynomial is defined in the following equation (3), for instance i = 1, in order not to contain X = 1 as a root:

$G(X) = (G - \alpha)(X - \alpha^2)\text{---}(X - \alpha^{d-1})$ (3)

where a symbol "d" indicates the minimum distance of the code.

SECOND QAM COMMUNICATION SYSTEM

In the above-described first QAM communication system 100, the 4-bit input signal has been coded as 1 symbol. Alternatively, in a second QAM communication system 200 shown in Fig. 4, an 8-bit input signal is coded as 1 symbol. The above-described third object of the present invention can be achieved by this second QAM communication system 200 employing the first basic idea and also the natural binary mapping method. That is, the second QAM communication system 200 has such a particular advantage that clock frequencies of error correction encoder/decoder can be selected to be lower than a modulating frequency of a quadrature amplitude modulator.

It should be noted that the same reference numerals shown in Fig. 3 will be employed as those for denoting the same or similar circuit elements shown in the following figures, and no further explanation thereof is made in the following descriptions.

In the second QAM communication system 200 shown in Fig. 4, digital signals inputted into 8 input

terminals 11-1 to 11-8 are supplied to 4 quadrature differential encoders 12-1 to 12-4 so as to be processed by a predetermined quadrature differential encoding process. The 8 differential-encoded 4-bit digital signals are furnished to two 4-to-8 bits parallel-to-parallel converters 30-1 and 30-2 thereby to obtain two pieces of 8-bit parallel data. These 8-bit parallel data are further supplied to 2 Reed-Solomon encoders 13-1 and 13-2 which are similar to the Reed-Solomon encoders employed in the first QAM communication system 100, so that these 8-bit parallel data are processed by a predetermined Reed-Solomon encoding process. Subsequently, these encoded 8-bit parallel data are input into a 256-QAM modulator 34 via two 8 to 4 bits parallel/parallel converters 32-1 and 32-2.

The 4-bit digital data which has been modulated in this 256-QAM modulator 34, is further supplied via the signal transmission channel 20 to a 256-QAM demodulator 36, whereby this 4-bit modulated digital data is demodulated therein. Thus, the demodulated 4-bit digital data is once converted into corresponding 8-bit parallel data by two sets of 4-to-8 bits parallel/parallel converters 38-1 and 38-2. Then, two pieces of 8-bit parallel data are decoded in the respective Reed-Solomon decoders 16-1 and 16-2. Thereafter, the decoded 8-bit parallel data are again converted into 8 pieces of 4-bit parallel data in two 8-to-4 bits parallel/parallel converters 39-1 and 39-2 and then are processed by a differential decoding process in 4 quadrature differential decoders 17-1 to 17-4, respectively. The resultant 8 pieces of digital signals are obtained from 8 output terminals 18-1 to 18-8.

Similarly, in accordance with the above-described second QAM communication system 200, the codes employed in these Reed-Solomon encoders 13-1 and 13-2 are $GF(2^8)$, and the generator polynomial thereof G(X) is not divisible by x-1. as the root. Since the Reed-Solomon encoders 13-1 and 13-2 perform the coding operations for the 8 bits digital signal as 1 symbol, there are provided the 4-to-8 bits parallel/parallel converting circuits 30-1 and 30-2 and the 8-to-4 bit parallel/parallel converting circuits 32-1 and 32-2 therebetween.

Assuming now that the generator polynomial G(X) of this code is expressed by:

$$G(X) = (X-\alpha)(X-\alpha^2)(X-\alpha^3)(X-\alpha^4)$$
$$= X^4 + \alpha^{76}X^3 + \alpha^{251}X^2 + \alpha^{81}X + \alpha^{10} \quad (4),$$

redundant of four symbols are added thereto every time 251 symbols are input into the Reed-Solomon encoders 13-1 and 13-2. An internal circuit of the respective Reed-Solomon encoders 13-1 and 13-2 is constructed as represented in, for instance, Fig. 5.

This Reed-Solomon encoder 13 per se shown in Fig. 5 is known in this field. In this Reed-Solomon encoder 13, four D flip-flops 42A to 42D, four multipliers 46A to 46D, and four adders 44A to 44D are mutually connected as represented in Fig. 5. These adders 44A to 44D perform the adding operation over GF (2^8), whereas the multipliers 48A to 48D perform the multiplication over $GF(2^8)$. As a result, the encoder 13 is constructed based upon the transparent Reed-Solomon code.

In the second QAM communication system 200 with the above-described circuit arrangement, the error less than 2 symbols occurring in the signal transmission channel 20 can be corrected by the Reed-Solomon decoders 16-1 and 16-2.

These Reed-Solomon decoders 16-1 and 16-2 may be realized by a circuit arrangement shown in Fig. 6.

Similarly, the Reed-Solomon decoder 16 itself represented in Fig. 6 is known in the art. This Reed-Solomon decoder 16 is constructed of a syndrome generator 50, an error location polynomial calculator 51, a Chien search circuit 52, an error value calculator 53, and an error correction circuit 54. Furthermore, a correction/detection controller 55 is employed to receive the output signals derived from the syndrome generator 50, error location polynomial calculator 51 and chien search circuit 52. Both the error location polynomial calculator 51 and error correction circuit 54 are under control of this correction/detection controller 55. To this error correction circuit 54, data which has been obtained by delaying the received 8-bit data is supplied.

In Fig. 7. there is shown an internal circuit arrangement of this syndrome generator 50. As apparent from Fig. 7, the syndrome generator 50 is so constructed by employing 4 sets of circuit arrangements each including an adder 50A, multiplier 50B, and D flip-flop 50C. These multipliers carry out the multiplication with α, α^2, α^3 and α^4 times. The syndrome output signals "S_1" to "S_4" are derived from these four lines and represented in the right-half portion of Fig. 7.

According to the above-described second preferred embodiment, the above-explained 2-symbol error correction can be realized at the coding rate of 98.4%. If the conventional binary BCH code for correcting two errors would be utilized under the same delay time condition, it will become (511,493) codes at the coding rate 96.5%. As a consequence, the frequency can be effectively utilized in the second QAM communication system 200. Moreover, if the conventional binary BCH code, or Lee error correction code is employed in the conventional QAM communication system, the clock frequencies of the encoders and

decoders cannot be lower than the modulating frequency. To the contrary, as described in the second preferred embodiment, in the case that 1 symbol of the Reed-Solomon code is allocated to 2 modulation symbols, the clock frequencies of the encoder 13 and decoder 16 can be selected to be a hall of the modulation velocity (modulating frequency). As previously stated, since the clock frequencies of the encoder 13 and decoder 16 can be lower than the modulalion velocity in accordance with the second preferred embodiment, there is a particular advantage in designing the error correction circuit of the high-speed data transmission system.

It should be noted that the generator polynomial with respect to the second QAM communication system 200 is not divisible by x-1.

As described in Fig. 12. many other modifications may be realized without departing from the first basic idea of the present invention, that is, the Reed-Solomon code is utilized while maintaining the transparency of the error correction code.

While the multi-level QAM communication system capable of having a transparent error correction with employing the first basic Idea of the present invention has been described, the frequency efficiency can be considerably increased as compared with that of the conventional QAM communication system. In other words, not only the frequency utilization efficiency can be improved, but also the total number ol the encoders/decoders can be reduced, in comparison with those of conventional QAM communication system with employing the binary BCH code.

Furthermore, not only the frequency utilization efficiency can be considerably improved, but also the reliability can be increased, as compared with the conventional multi-level Lee error correction code. This system has a particular merit when utilized in a mobile communication with a fading phenomenon. According to the first basic idea of the present invention, it can provide a multi-level error correction code capable of maintaining the transparency even when the differential encoding/decoding operations are performed. That is, the transparency can be maintained by the following methods. Namely, in case that the signal point mapping is the natural code mapping, such a Reed-Solomon code is employed that (X - 1) is not included in the generator polynomial as the factor.

THIRD QAM COMMUNICATION SYSTEM

Referring now to Fig. 8, a 256-QAM communication system 300 employing the second basic idea, according to a third preferred embodiment of the present Invention, will be described.

Fig. 8 is a schematic block diagram of the 256-QAM communication system 300.

In Fig. 8 at a signal transmission side, as viewed in the left side of this drawing, 8 digital signal series are inputted via 8 input terminals 11-1 to 11-8. The 2 bits digital signals supplied from two input terminals 11-1 and 11-2 are furnished to a quadrature differential encoder 32 so as to be differential-encoded therein. Then, the resultant 2 bits differentially-encoded signals are separately supplied to first and second encoders 33 and 34 having the same function with each other, whereby 2 bits encoded data are obtained therefrom. On the other hand, 6 bits digital signals inputted from the remaining 6 input terminals 11-3 to 11-8 are supplied to a third encoder 35 thereby to obtain 6 encoded digital signals. It should be noted that the codes employed in the first and second encoders 33 and 34 are Identical to each other, and correspond to such an error correction code having all 1 vectors as the code words. The code employed in the third encoder 35 may be selected to be an arbitrary code. When, for instance, a Reed-Solomon code on GF(2^6) capable of correcting a symbol error is utilized as this code for the third encoder 35, there is a particular effect since the correction capability is great with respect to redundancy.

The 8 bits digital signals derived from the first to third encoders 33, 34 and 35 are converted by a 256-QAM modulator 36 into signal waveforms corresponding to signal points which have been mapped based upon the quadrant symmetry mapping, and thereafter output to a signal transmission channel 20.

On the other hand, at a signal reception side, 8 bits digital signals which have been demodulated by a 256-QAM demodulator 38 are error-corrected by first to third decoders 39, 40 and 41 which correspond to the first to third encoders 33, 34 and 35. The 2 bits digital signals output from the first and second decoders 39 and 40 are differential-decoded in a quadrature differential decoder 42. Thus, the two signal series derived from the differential decoder 42 are outputted from output terminal 18-1 and 18-2 , whereas the remaining signal series derived from the third decoder 41 are outputted from other output terminals 18-3 to 18-8.

In the 256 QAM communication system 300 of the quadrant symmetry mapping according to the third preferred embodiment, the error correction coding operations are separately performed with respect to the

combinations between the signal series for determining the quadrant of the phase plane, and other signal series, so that the transparent coding operation can be realized with respect to the phase rotations occurring between the input digital signals and output digital signals.

As a consequence, when the combinations of the signal series which have no relation to determine the quadrant of the phase plane are encoded, since a plurality of signal series can be encoded as inputs, the resultant circuit arrangement can be made small, as compared with such a case that all of the signal series are independently encoded.

FOURTH QAM COMMUNICATION SYSTEM

Furthermore, a 256-QAM communication system 400 utilizing the second basic idea, according to a fourth preferred embodiment of the present invention, will now be described with reference to Fig. 9.

That is, Fig. 9 is a schematic block diagram for representing the fourth 256-QAM communication system 400 of the quadrant symmetry mapping.

At a signal transmission side, as viewed in a left side of Fig. 9, 8 digital signal series are inputted into this system 400 via 8 input terminals 11-1 to 11-8. The 2 bits digital signals inputted from the two input terminals 11-1 and 11-2 are supplied into a differential encoder 45 so as to be differentially-encoded. The resultant 2 bits differentially-encoded signals are separately supplied to a first encoder 46 and a second encoder 47 having the same function as that of the first encoder 46, whereby 2 bits encoded digital signals are produced therefrom respectively. Another 2-bit digital signal combination inputted from the subsequent two input terminals 11-3 and 11-4 is furnished to a third encoder 48 so as to be encoded. Similarly, each of 2-bit digital signal combinations which are inputted from two input terminals 11-5 and 11-6, and also two input terminals 11-7 and 11-8 respectively, is supplied to fourth and fifth encoders 49 and 50 respectively for signal encoding purposes.

It should be noted that the codes employed in the first and second encoders 46 and 47 commonly connected to the differential encoder 45 are the same with each other, and therefore correspond to such error correcting codes having all 1 vectors as code words. Also, error correcting codes utilized in the third, fourth, and fifth encoders 48, 49 and 50 may be selected to be arbitrary, for instance, an error correcting code having an error correcting capability in accordance with a bit error rate of signal series.

Then, the 8 bits digital signals outputted from the first to fifth encoders 46, 47, 48, 49 and 50 are converted in a 256-QAM modulator 51 into signal waveforms corresponding to the signal points which have been mapped based upon the quadrant symmetry mapping.

At a signal reception side of the fourth 256-QAM communication system 400, the 8 bits digital signals which have been demodulated in a 256-QAM demodulator 53, are error-corrected by first to fifth decoders 54, 55, 56, 57 and 58. The 2 bits decoded digital signals derived from the first and second decoders 54 and 55 are differential-decoded by a quadrature differential decoder 59 corresponding to the above-described quadrature differential encoder 54, so that 2 bit differentially-encoded signals are outputted from this quadrature differential decoder 59 to two output terminals 18-1 and 18-2, respectively. The remaining 2-bit decoded digital signal combinations are directly outputted from the third to fifth decoders 56 to 58 to other output terminals 18-3 to 18-8.

As a result, also in the fourth preferred embodiment, when the combinations of the signal series which has no relation to determine the quadrant of the phase plane are encoded, the coding operation can be done with each of the 2-input signal combinations. Consequently, the overall system can be made small, as compared with another system in which all of the signal series are independently encoded.

There is another particular advantage that since the redundancy of the error correcting codes can be determined based upon the bit error rate for the receptions of the respective signal series, the coding operations can be effectively executed with smaller redundancy.

FIFTH QAM COMMUNICATION SYSTEM

In Fig. 10, there is shown a 256-QAM communication system 500 by the Gray code, according to a fifth preferred embodiment of the present invention. It should be noted that this fifth QAM communication system 500 similarly employs the second basic idea of the invention.

At a signal transmission side of the fifth 256-QAM communication system 500, 8 digital signal series are

inputted from 8 input terminals 11-1 to 11-8, and supplied to a differential encoder 62. In the differential encoder 62, these input digital signals are differentially-encoded, and two encoded signal series I_1 and Q_2 are supplied to first and second encoders 63 and 64 having the same functions with each other for the encoding purposes. Three encoded signal series I_2, I_3 and I_4 are supplied to a third encoder 65 for the encoding purposes. Similarly, three encoded signal series Q_2, Q_3 and Q_4 are encoded in a fourth encoder 66. It should be noted that the codes employed in the first and second encoders 63 and 64 are the same with each other, and correspond to an error correcting word having all 1 vectors as code words. Further, error correcting codes utilized in the third and fourth encoders 65 and 66 are the same with each other and are arbitrary codes. The 8 bits digital signals derived from the first to fourth encoders 63, 64, 65 and 66 are converted by a 256-QAM modulator 67 into waveforms corresponding to signal points which have been mapped based upon the above-described Gray code, and the Gray-coded digital signals are output therefrom to the signal transmission channel 20.

On the other hand, at a signal reception side of this 256-QAM communication system 500, the 8 bits digital signals which have been demodulated by a 256-QAM demodulator 89 are error-corrected by first to fourth decoders 70, 71, 72 and 73, and thereafter supplied to a differential decoder 74. Thus, these 8 bits digital signals derived from the first to fourth decoders 70 to 73 are differential-decoded. The resultant 8 bits differential-decoded digital signals are outputted from 8 output terminals 18-1 to 18-8, respectively.

When the combinations of the signal series I_2 through Q_4 which have no relation to determine the quadrant of the phase plane are encoded, since each of 3-input signal series I_2, I_3, I_4 and Q_2, Q_3, Q_4 can be encoded in the third and fourth encoders 65 and 66, the overall system 500 can be made small, as compared with another system in which all of the signal series are separately coded.

As previously described in detail, in accordance with the respective third to fifth QAM communication systems, when the combinations of the signal series having no relation to determine the quadrant of the phase plane are encoded, since a plurality of signal series are encoded, the entire circuit arrangement of the QAM communication system can be simply constructed, as compared with such a QAM communication system that all of the input signal series are separately encoded.

SIXTH QAM COMMUNICATION SYSTEM

In Fig. 11 there is shown a 64 QAM communication system 600 employing the third basic idea, according to a sixth preferred embodiment of the present invention.

As apparent from Fig. 11, a major error control circuit portion of the sixth QAM communication system 600 is substantially identical to that of the second QAM communication system 200. Accordingly, the same reference numerals shown in Fig. 4 will be employed as those for denoting the same or similar circuit elements in the following drawings, and no further explanation thereof will be made in the following descriptions.

That is, both a 64-QAM modulator 80 and a 64-QAM demodulator 82 are newly employed in this sixth preferred embodiment.

An operation of the 64-QAM communication system 600 will now be described.

It should be noted that signal information of two signal series 11-1 and 11-2 outputted from the quadrature differential encoder 12 determine the quadrant of the signal phase plane in the 64-QAM modulator 80. In case of the quadrant symmetry mapping, 4 signal series 11-3 through 11-6 other than the above-described 2 signal series 11-1 and 11-2 receive no adverse influence of the signal phase rotation, so that an arbitrary code may be employed. When a GF (2^8) Reed-Solomon code is employed in these signal series, an error correcting method with a high coding rate of 99.0 % may be utilized. In addition, the clock frequency of the error correcting circuits can be made half of the modulation frequency. This error correcting method shown in Fig. 11 enables the transparency to be established with respect to the signal phase rotation.

Further, if the signal point mapping is the quadrant symmetry mapping, and the above-described two signal series 11-1 and 11-2 which are not encoded by the Reed-Solomon code are independently encoded by transparent error correcting codes, this modified QAM communication system may become a transparent error correcting system without receiving any adverse influence caused by the signal phase rotation.

SEVENTH QAM COMMUNICATION SYSTEM

Fig. 12 is a schematic block diagram of a 64-QAM communication system 700 according to a seventh preferred embodiment of the present invention, with employment of the first basic idea and the natural binary mapping method as shown in Fig. 14.

In the seventh 64-QAM communication system 700, there are newly provided 6 pairs of 1-bit serial data to 8-bit parallel data converters 84-A to 84-F and 86-A to 86-F, and also 6 pairs of 8-bit parallel data to 1-bit serial data converters 85-A to 85-F and 88-A to 88-F. The main feature of the seventh preferred embodiment is that all of 6 input digital signals supplied from the input terminals 11-1 to 11-8 are independently processed with the Reed-Solomon encoders 83-1 to 83-5 and Reed-Solomon decoders 87-1 to 87-5. In case of natural binary mapping, all Reed-Solomon code should be transparent and identical.

That is to say, when as shown in Fig. 12, the Reed-solomon codes are applied to all ol the signal series, it can be achieved such an error correcting system capable of correcting a burst error. When the above-described 2-symbol (255, 251) error correction Reed-Solomon code on GF (2^8) is employed as the Reed-Solomon code, there is a particular advantage that a single burst error having a burst length of less than 9 bits can be corrected. In addition, the clock frequency of error control encoders/decoders can be made as low as 1/8 times the modulation frequency.

As previously described in detail, in accordance with the respective third to seventh QAM communication systems, when the combinations of the signal series having no relation to determine the quadrant of the phase plane are encoded, since a plurality of signal series are encoded, the entire circuit arrangement of the QAM communication system can be simply constructed, as compared with such a QAM communication system that all of the input signal series are separately encoded.

EIGHTH QAM COMMUNICATION SYSTEM

Furthermore, in accordance with the present invention, many other modifications may be achieved other than the above-described first to seventh preferred embodiments 100 to 700 with employment of the first and second basic Ideas.

For instance, in case that the phase ambiguity or phase ambiguity of the reproduced carrier wave is removed by periodically transmitting the known signal patterns, there is no need to separately employ the error correction encoder/decoder (i.e., quadrature differential encoder/decoder) with respect to the I-axis and Q-axis, and "n" pieces of signal series for determining the signal point mapping are considered as a symbol which will be coded as a Reed-Solomon code.

A 256-QAM communication system 800 utilizing the above-described idea, according to an eighth preferred embodiment of the present invention, will now be described with reference to Fig. 13. As apparent from the circuit arrangement shown in Fig. 13, no quadrature differential encoder/decoder is employed in the eighth 256-QAM communication system 800. Instead of the quadrature differential encoder/decoder, a unique word adder 820 for producing the above-described known signal pattern is connected to the output terminal of the Read-Solomon encoder 810, and also a unique word detector 850 is connected to the input terminal of the Read-Solomon decoder 860.

After 8 series of 255 bits signal are transmitted, a unique word constructed of 1 symbol (8 bits) is transmitted from the unique word adder 820 in the eight 256-QAM communication system 800. The quadrant phase information is added to the Read-Solomon encoded input signals based upon this unique word, since the unique word has a function to correct phase shifts. Thus, the data to which the unique word has been added and which has been modulated in the 256-QAM modulator 830 and thereafter demodulated in the 256-QAM demodulator 840, contains no longer such phase ambiguity, or phase ambiguity. Subsequently, the unique word is detected by the unique word detector 850 from this data so as to obtain absolute phases with respect to the input signals. As a result, the phase ambiguity has been eliminated from the resultant 8-bit data derived from the Reed-Solomon decoder 880 based upon the detected unique word.

There are the following advantages of the eighth 256-QAM communication system 800.

In case of the 256-QAM system shown in Fig. 13, since n = 8, then a code of GF (2^8) having a length of 255 can be utilized. When a two-symbol correction code is employed similar to the above-described second preferred embodiment,. the resultant coding rate can become 98.4%. However, the conventional binary BCH code to perform such a 2 error correction under the same delay time becomes a code (255,239) at a coding rate of 93.7%. As a consequence, the frequency utilization efficiency according to the present invention can be improved. Furthermore, although the total number of the respective encoders and decoders is 8 in case of the conventional binary BCH code, only 1 Read-Solomon and 1 Read-Solomon

decoder are required in the eighth preferred embodiment. Consequently, there is another particular advantage that a scale of the entire circuit arrangement can be reduced.

As is apparent from the foregoings, the Reed-Solomon encoder 810 and decoder 860 may be provided inside the unique word adder 820 and detector 850. Furthermore, both the unique word adder and detector may be omitted from the eighth QAM communication system 800, and alternatively, the Reed-Solomon encoder/decoder may employ an absolute phase detecting function.

Reference signs in the claims are intended for better understanding and shall not limit the scope.

Claims

1. A multi-level QAM (quadrature amplitude modulation) communication system (100;200;700) utilizing n bits ("n" being an integer) QAM signal having 2^n signal points, comprising:
differential encoder/decoder means (12;17) for differentially encoding/decoding n pieces of input digital signal series to resolve phase ambiguity contained in the differentially encoded input signal series;
error correction means including a Reed-Solomon encoder (13;83) and a Reed-Solomon decoder (16;87), provided inside said differential encoder/decoder means (12;17) along a signal processing path of said input digital signal series, for error-control-encoding/decoding said n pieces of differentially-coded signal series by utilizing at least one of said digital signal series to correct errors with employment of Reed-Solomon codes; and,
QAM modulator/demodulator means (14;15;34;36;80;82) for QAM-modulating/demodulating n pieces of error-control-coded signal series so as to produce 2^n QAM signals.

2. A multi-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 1, wherein said error correction means (13;16) separately performs encoding/decoding operations with employment of the same Reed-Solomon code with respect to two orthogonalized channels (I;Q).

3. A multi-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 1, wherein said differential encoder/decoder means (12;17) performs a differential coding operation under such a condition that a generator polynomial of the Reed-Solomon code is not divisible by X-1, said generator polynomial being given as

$$G(X) = \prod_{i=r}^{d-2+r} (X - \alpha^i),$$

where "i" is an integer, "α" is a primitive element of Galois field, and "d" is the minimum distance of the Reed-Solomon code.

4. A multi-level AQM (quadrature amplitude modulation) communication system as claimed in Claim 1, wherein a total number of said error correction means is smaller than a bit number of said QAM signal.

5. A multi-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 1, wherein said signal point mapping is a natural binary mapping.

6. A multi-level QAM (quadrature amplitude modulation) communication system (300;400;500) for producing n bits ("n" being an integer) QAM signal having 2^n signal points from n pieces of input digital signal series, comprising:
differential encoder/decoder means (32;45;62;42;59;74) for differentially encoding/decoding at least two pieces of signal series for determining a quadrant of a phase plane for a signal point mapping among said n pieces of input digital signal series so as to produce at least two pieces of differentially-coded signal series;
first error correction means (33,34;39,40:46,47;54,55;63,64;70,71) provided inside said differential encoder/decoder means (12;32;74;17) along a signal processing path of said input digital signal series, for error correction so as to produce at least two pieces of first error-control-coded signal series;
second error correction means (35;48,49,50;65,66:41;56,57,58;72,78) provided directly to receive remaining pieces of input digital signal series so as to produce second error-control-coded signal series; and,
QAM modulator/demodulator means (36;51:38,53) for QAM-modulating/demodulating both said first error-control-coded signal series and second error-control-coded signal series so as to output the n bits QAM signal, whereby phase ambiguity contained in the differentially encoded input signal series is solved.

7. A multi-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 6, wherein said second error correction means is constructed of a plurality of error correction encoders (48;49;50:65;66) and a plurality of error correction decoders (56;57;58:72;73).

8. A multi-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 6, wherein said second error correction means employs a nonbinary error correcting code.

9. A multi-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 6, wherein said signal point mapping is a quadrant symmetry mapping.

10. A multi-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 6, wherein said signal point mapping is a Gray code mapping.

11. A multi-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 6, wherein said differential encoder/decoder means (32,42;45,59;62,74) performs a differential coding operation under such a condition that a generator polynomial of an error control code is not divisible by X-1, said generator polynomial being given as $G(X) = LCM\ \{m_r(X), m_{r+1}(X), \ldots, m_{d-2+r}(X)\}$, where "r" is an integer, $m_i(X)$ is the minimum function of α_i, "α" is a primitive element of Galois field, and "d" is the minimum distance of the error control code.

12. A mulli-level QAM (quadrature amplitude modulation) communication system as claimed in Claim 6, wherein a total number of said first and second error correction means is smaller than a bit number of said QAM signal.

13. A multi-level QAM communication system as claimed in Claim 6, wherein said first error correction means includes:
two sets of error correction code encoders/decoders (33,39:46,54;64:71) for each encoding and docoding one differentially-coded signal series derived from said differential encoder/decoder means so as to finally produce said first error-control-coded signal series, each of said error correction codes being the same code with each other.

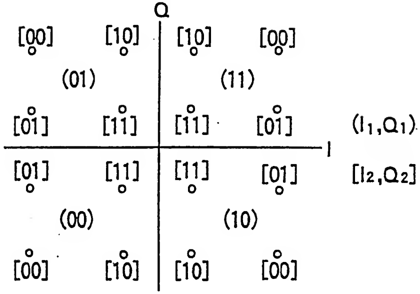
14. A multi-level QAM (quadrature amplitude modulation) communication system (600:800) for producing 2^n pieces of n-bit ("n" being an integer) QAM signal points from n pieces of input signal series, comprising:
encoder/decoder means (12,13;17,16:810,860) utilizing a Reed-Solomon code of Galols field $GF(2^l)$, for error-correction-encoding/decoding h pieces of input signal series ("h" being a positive integer smaller than or equal to "n" and "l" being larger than or equal to "h"); and
QAM-modulating/demodulating means (80,82;830,840) positioned inside said encoder/decoder means along a signal processing path of said n pieces of input signal series, for QAM-modulating/demodulating said n pieces of input signal series containing said h pieces of Reed-Solomon encoded input signal series so as to resolve phase ambiguity contained in said h pieces of Reed-Solomon encoded signals.

15. A multi-level QAM (quadrature amplitude modulation) communication system (600:800) as claimed in Claim 14, wherein said "l" is equal to A x h, where "A" is a positive integer.

16. A multi-level QAM (quadrature amplitude modulation) communicalion system as claimed in Claim 14, further comprising:
unique word adding means (820) for adding a unique word representative of quadrant phase information on said input signal series to said h pieces of Reed-Solomon encoded signal series; and,
unique word detecting means (850) for detecting said unique word contained in said h pieces of Reed-Solomon encoded signal series in order to obtain absolute phases with respect to said input digital signals.

17. A multi-level QAM (quadrature amplitude modulation) communication system (600) as claimed in Claim 14, further comprising:
quadrature differential encoder/decoder means (12;17) for differentially encoding/decoding (n-h) pieces of input digital signal series to produce (n-h) pieces of differentially-coded signal series.

PRIOR ART
FIG.1
(GRAY MAPPING)



PRIOR ART
FIG.2
(QUADRANT SYMMETRY MAPPING)

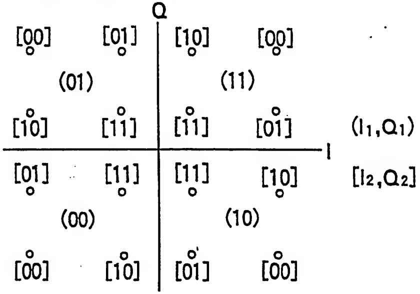


FIG.3

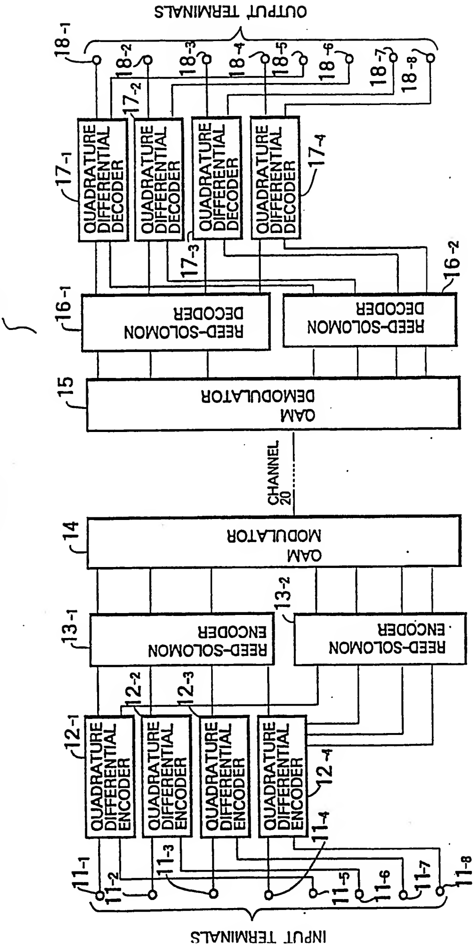

100
OUTPUT TERMINALS
18-1
18-2
18-3
18-4
18-5
18-6
18-7
18-8
17-1
17-2
17-3
17-4
QUADRATURE DIFFERENTIAL DECODER
QUADRATURE DIFFERENTIAL DECODER
QUADRATURE DIFFERENTIAL DECODER
QUADRATURE DIFFERENTIAL DECODER
16-1
16-2
REED-SOLOMON DECODER
REED-SOLOMON DECODER
15
QAM DEMODULATOR
CHANNEL 20
14
QAM MODULATOR
13-1
13-2
REED-SOLOMON ENCODER
REED-SOLOMON ENCODER
12-1
12-2
12-3
12-4
QUADRATURE DIFFERENTIAL ENCODER
QUADRATURE DIFFERENTIAL ENCODER
QUADRATURE DIFFERENTIAL ENCODER
QUADRATURE DIFFERENTIAL ENCODER
11-1
11-2
11-3
11-4
11-5
11-6
11-7
11-8
INPUT TERMINALS

FIG.4

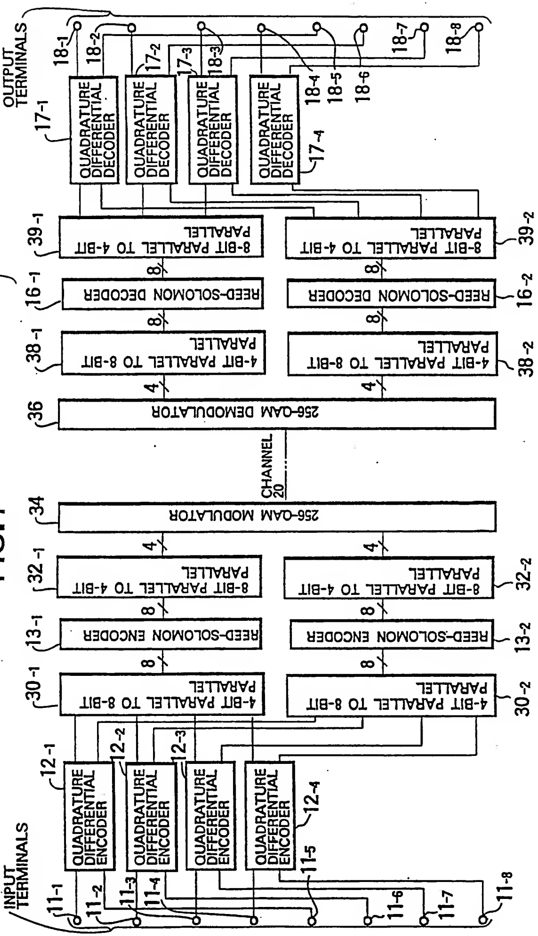

200
INPUT TERMINALS
11-1
11-2
11-3
11-4
11-5
11-6
11-7
11-8
QUADRATURE DIFFERENTIAL ENCODER
12-1
12-2
12-3
12-4
4-BIT PARALLEL TO 8-BIT PARALLEL
30-1
30-2
8
REED-SOLOMON ENCODER
13-1
13-2
8-BIT PARALLEL TO 4-BIT PARALLEL
32-1
32-2
4
256-QAM MODULATOR
34
CHANNEL 20
256-QAM DEMODULATOR
36
4-BIT PARALLEL TO 8-BIT PARALLEL
38-1
38-2
REED-SOLOMON DECODER
16-1
16-2
8-BIT PARALLEL TO 4-BIT PARALLEL
39-1
39-2
QUADRATURE DIFFERENTIAL DECODER
17-1
17-2
17-3
17-4
OUTPUT TERMINALS
18-1
18-2
18-3
18-4
18-5
18-6
18-7
18-8

FIG.5

REED-SOLOMON ENCODER 13

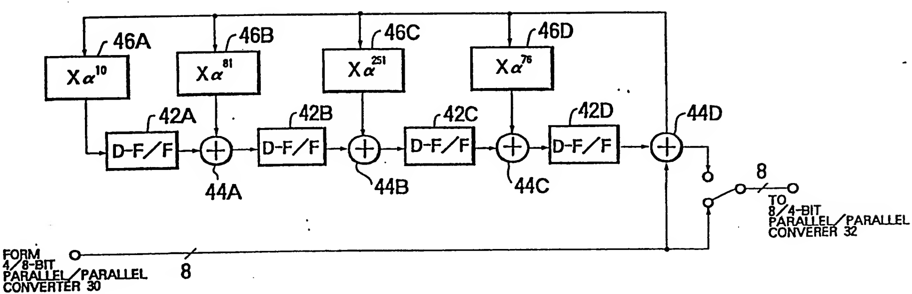


FIG.6

REED-SOLOMON DECODER 16

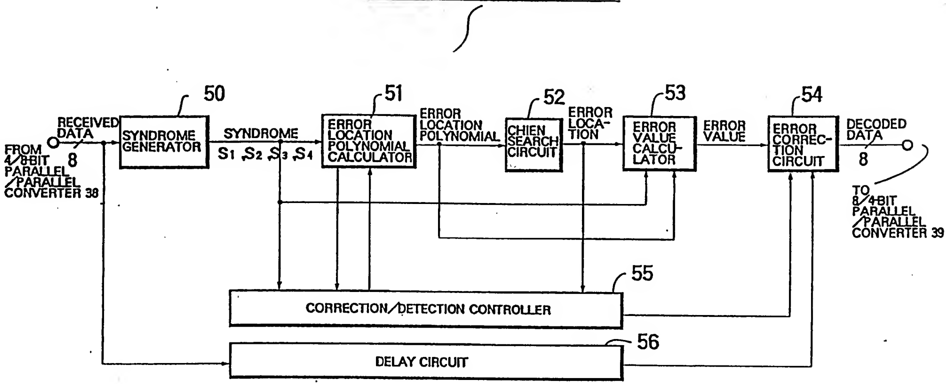


FIG.7

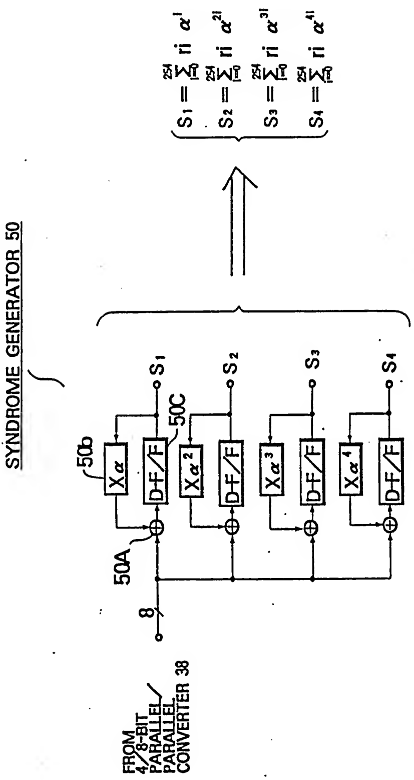

SYNDROME GENERATOR 50
FROM 4/8-BIT PARALLEL/PARALLEL CONVERTER 38
8
50A
50b
50C
Xα
Xα²
Xα³
Xα⁴
D-F/F
S1
S2
S3
S4
S1 = Σ ri α^i
S2 = Σ ri α^2i
S3 = Σ ri α^3i
S4 = Σ ri α^4i

FIG.8

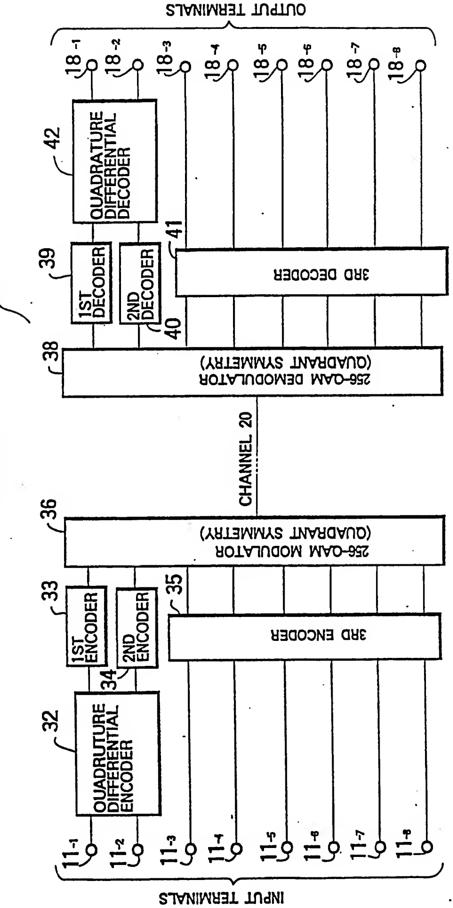
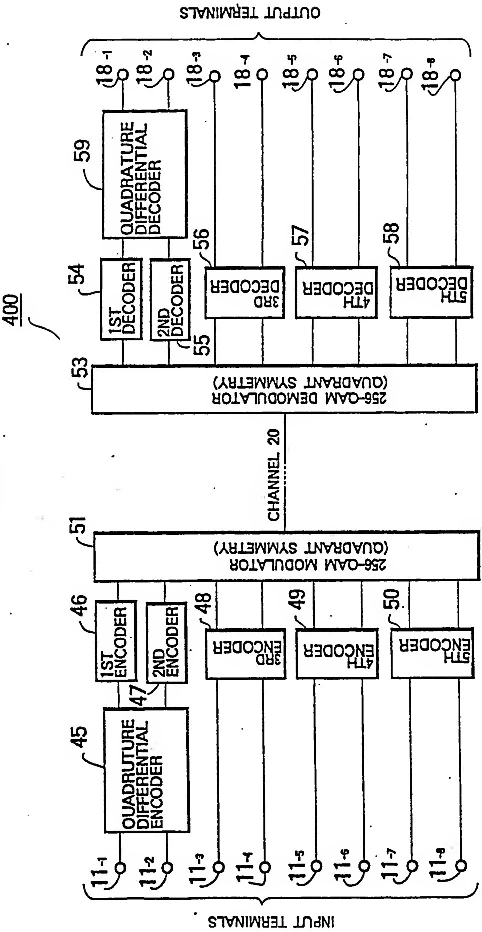

300
OUTPUT TERMINALS
18-1
18-2
18-3
18-4
18-5
18-6
18-7
18-8
42
QUADRATURE DIFFERENTIAL DECODER
39
1ST DECODER
2ND DECODER
40
41
3RD DECODER
38
256-QAM DEMODULATOR (QUADRANT SYMMETRY)
CHANNEL 20
36
256-QAM MODULATOR (QUADRANT SYMMETRY)
33
1ST ENCODER
34
2ND ENCODER
35
3RD ENCODER
32
QUADRUTURE DIFFERENTIAL ENCODER
11-1
11-2
11-3
11-4
11-5
11-6
11-7
11-8
INPUT TERMINALS

FIG.9


400
OUTPUT TERMINALS
18-1
18-2
18-3
18-4
18-5
18-6
18-7
18-8
59
QUADRATURE DIFFERENTIAL DECODER
54
1ST DECODER
2ND DECODER
55
56
3RD DECODER
57
4TH DECODER
58
5TH DECODER
53
256-QAM DEMODULATOR (QUADRANT SYMMETRY)
CHANNEL 20
51
256-QAM MODULATOR (QUADRANT SYMMETRY)
46
1ST ENCODER
47
2ND ENCODER
48
3RD ENCODER
49
4TH ENCODER
50
5TH ENCODER
45
QUADRUTURE DIFFERENTIAL ENCODER
11-1
11-2
11-3
11-4
11-5
11-6
11-7
11-8
INPUT TERMINALS

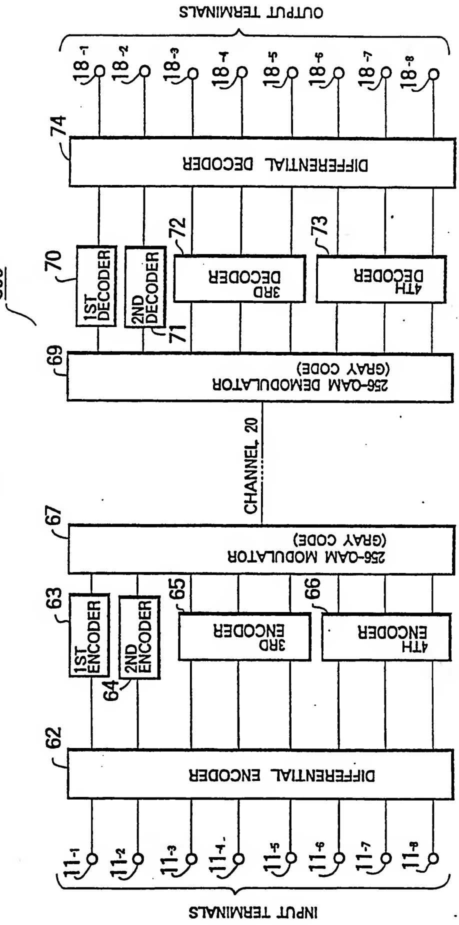

500
OUTPUT TERMINALS
18-1
18-2
18-3
18-4
18-5
18-6
18-7
18-8
74
DIFFERENTIAL DECODER
70
1ST DECODER
2ND DECODER
71
72
3RD DECODER
73
4TH DECODER
69
256-QAM DEMODULATOR (GRAY CODE)
CHANNEL 20
67
256-QAM MODULATOR (GRAY CODE)
63
1ST ENCODER
64
2ND ENCODER
65
3RD ENCODER
66
4TH ENCODER
62
DIFFERENTIAL ENCODER
11-1
11-2
11-3
11-4
11-5
11-6
11-7
11-8
INPUT TERMINALS

FIG.10

FIG.11

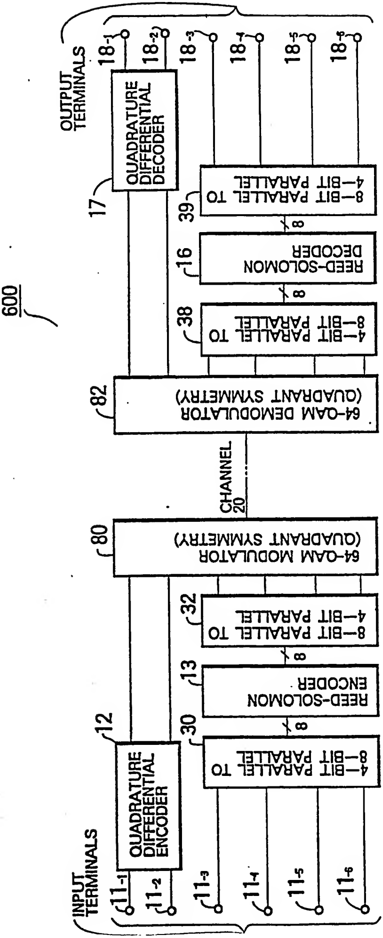

600
INPUT TERMINALS
11-1
11-2
11-3
11-4
11-5
11-6
12
QUADRATURE DIFFERENTIAL ENCODER
30
4—BIT PARALLEL TO 8—BIT PARALLEL
8
13
REED-SOLOMON ENCODER
8
32
8—BIT PARALLEL TO 4—BIT PARALLEL
80
64-QAM MODULATOR (QUADRANT SYMMETRY)
CHANNEL 20
82
64-QAM DEMODULATOR (QUADRANT SYMMETRY)
38
4—BIT PARALLEL TO 8—BIT PARALLEL
8
16
REED-SOLOMON DECODER
8
39
8—BIT PARALLEL TO 4—BIT PARALLEL
17
QUADRATURE DIFFERENTIAL DECODER
OUTPUT TERMINALS
18-1
18-2
18-3
18-4
18-5
18-6

FIG.12

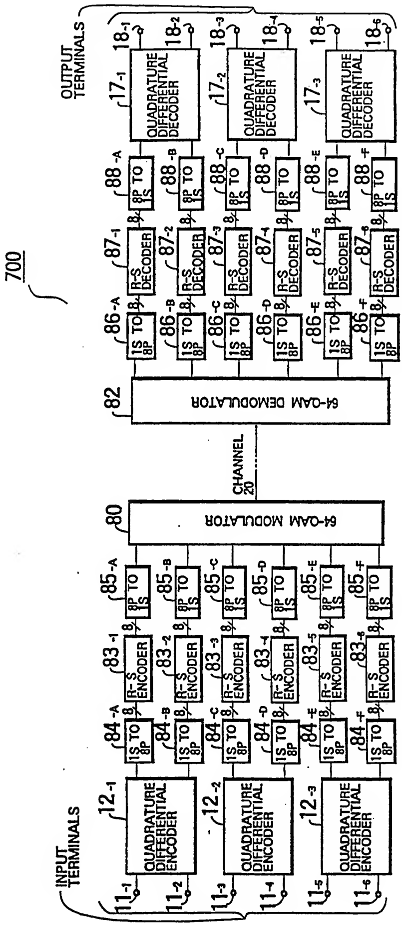
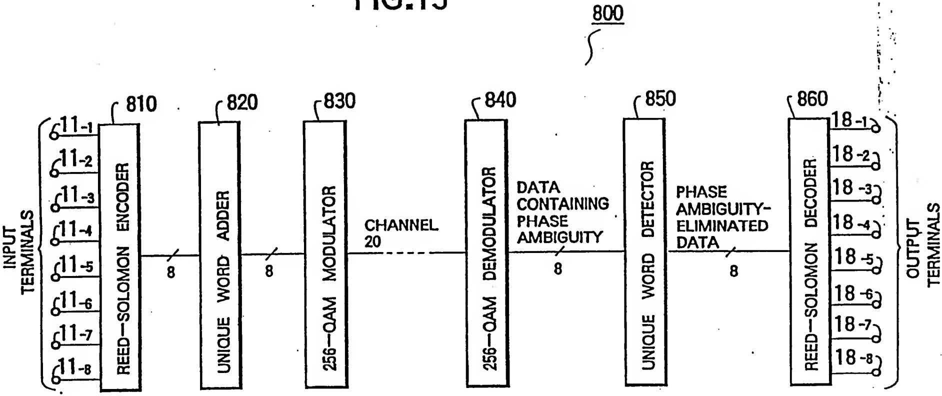

700
INPUT TERMINALS
11-1
11-2
11-3
11-4
11-5
11-6
12-1
12-2
12-3
QUADRATURE DIFFERENTIAL ENCODER
84-A
84-B
84-C
84-D
84-E
84-F
1S TO 8P
8
83-1
83-2
83-3
83-4
83-5
83-6
R-S ENCODER
85-A
85-B
85-C
85-D
85-E
85-F
8P TO 1S
80
64-QAM MODULATOR
CHANNEL 20
82
64-QAM DEMODULATOR
86-A
86-B
86-C
86-D
86-E
86-F
1S TO 8P
87-1
87-2
87-3
87-4
87-5
87-6
R-S DECODER
88-A
88-B
88-C
88-D
88-E
88-F
8P TO 1S
17-1
17-2
17-3
QUADRATURE DIFFERENTIAL DECODER
18-1
18-2
18-3
18-4
18-5
18-6
OUTPUT TERMINALS

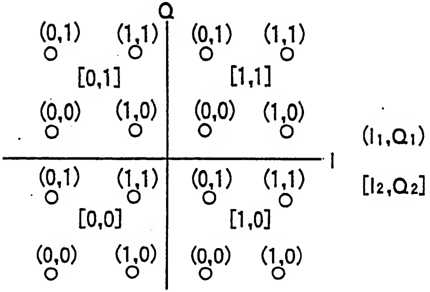
FIG.13


800
INPUT TERMINALS
11-1
11-2
11-3
11-4
11-5
11-6
11-7
11-8
810
REED—SOLOMON ENCODER
8
820
UNIQUE WORD ADDER
8
830
256—QAM MODULATOR
CHANNEL 20
840
256—QAM DEMODULATOR
DATA CONTAINING PHASE AMBIGUITY
8
850
UNIQUE WORD DETECTOR
PHASE AMBIGUITY-ELIMINATED DATA
8
860
REED—SOLOMON DECODER
18-1
18-2
18-3
18-4
18-5
18-6
18-7
18-8
OUTPUT TERMINALS

PRIOR ART

FIG.14

(NATURAL BINARY MAPPING)



PRIOR ART

FIG.15A NATURAL BINARY MAPPING

PHASE REFERENCE ERROR	RECEIVED DATA	
	$I_1\ I_2 — I_{n/2}$	$Q_1\ Q_2 — Q_{n/2}$
0	$x_1\ x_2 — x_{n/2}$	$y_1\ y_2 — y_{n/2}$
$\pi/2$	$y_1\ y_2 — y_{n/2}$	$\bar{x}_1\ \bar{x}_2 — \bar{x}_{n/2}$
π	$\bar{x}_1\ \bar{x}_2 — \bar{x}_{n/2}$	$\bar{y}_1\ \bar{y}_2 — \bar{y}_{n/2}$
$3\pi/2$	$\bar{y}_1\ \bar{y}_2 — \bar{y}_{n/2}$	$x_1\ x_2 — x_{n/2}$

PRIOR ART

FIG.15B GRAY MAPPING

PHASE REFERENCE ERROR	RECEIVED DATA	
	$I_1 \cdot I_2 — I_{n/2}$	$Q_1\ Q_2 — Q_{n/2}$
0	$x_1\ x_2 — x_{n/2}$	$y_1\ y_2 — y_{n/2}$
$\pi/2$	$y_1\ y_2 — y_{n/2}$	$\bar{x}_1\ x_2 — x_{n/2}$
π	$\bar{x}_1\ x_2 — x_{n/2}$	$\bar{y}_1\ y_2 — y_{n/2}$
$3\pi/2$	$\bar{y}_1\ y_2 — y_{n/2}$	$x_1\ x_2 — x_{n/2}$

PRIOR ART

FIG.15C QUADRANT SYMMETRY MAPPING

PHASE REFERENCE ERROR	RECEIVED DATA	
	$I_1\ I_2 — I_{n/2}$	$Q_1\ Q_2 — Q_{n/2}$
0	$x_1\ x_2 — x_{n/2}$	$y_1\ y_2 — y_{n/2}$
$\pi/2$	$y_1\ x_2 — x_{n/2}$	$\bar{x}_1\ y_2 — y_{n/2}$
π	$\bar{x}_1\ x_2 — x_{n/2}$	$\bar{y}_1\ y_2 — y_{n/2}$
$3\pi/2$	$\bar{y}_1\ x_2 — x_{n/2}$	$x_1\ y_2 — y_{n/2}$

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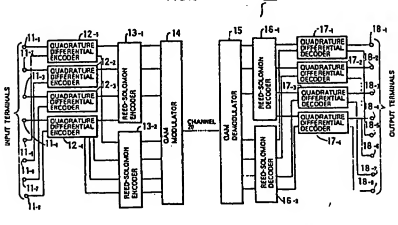
(54) Quadrature amplitude modulation communication system with transparent error correction.

(57) In a multi-level QAM communication system, Reed-Solomon encoders and Reed-Solomon decoders are employed for error correction purposes. The phase ambiguity of the received signal is eliminated with differential coding. The multi level QAM communication system (100) utilizing n bits ("n" being an integer) QAM signal having 2^n signal points, comprises: a quadrature differential encoder/decoder unit (12;17) for differentially encoding/decoding n pieces of input digital signal series to produce n pieces of differentially coded signal series; an error correction unit including a Reed-Solomon encoder (13) and a Reed-Solomon decoder (16), provided inside the quadrature differential encoder/decoder unit (12;17) along a signal processing path of the input digital signal series, for error-correcting the n pieces of differentially-coded signal series by utilizing at least one of the digital signal series with employment of a Reed-Solomon code; and, a QAM modulator/demodulator unit (14;15;34;35) for QAM-modulating/demodulating n pieces of error-corrected signal series so as to produce 2^n QAM signals.

FIG.3



European Patent Office

EUROPEAN SEARCH REPORT

Application Number

EP 90107036.7

DOCUMENTS CONSIDERED TO BE RELEVANT			
Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.5)
A	EP - A2 - 0 185 496 (AMERICAN TELEPHONE AND TELEGRAPH) * Abstract; page 3, line 11 - page 5, line 31; page 9, line 16 - page 12, line 34; fig: 1,3 * --	1,6	H 04 L 27/34 H 03 M 13/00 H 03 M 7/36 H 03 M 7/40 H 03 D 3/02
A	EP - A2 - 0 134 101 (AMERICAN TELEPHONE AND TELEGRAPH) * Page 1, line 6 - page 4, line 19 * --	1,6	
D,A	IEEE INTERNATIONAL CONFERENCE ON COMMUNICATIONS 1986, June 22-25, 1986, Sheraton Centre, Toronto Canada ICC '86 "Integrating the Word through communications" Conference Record, vol. 3 of 3 Y. YOSHIDA et al. "6GHZ 140MBPS Digital Radio Repeater with 256QAM Modulation" pages 1482-1486 * Totality * --	1,6	TECHNICAL FIELDS SEARCHED (Int. Cl.5) H 04 L H 03 M H 03 D
A	EP - A2 - 0 296 828 (SONY) --		
A	EP - A1 - 0 133 137 (ETABLISSEMENT PUBLIC DE DIFFUSION) ----		

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
VIENNA	10-12-1990	HAJOS

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
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T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

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(54) **Multiplexed coded modulation with unequal error protection.**

(57) Unequal error protection is provided for an HDTV signal (101) by separately coding (in 120,130) each one of the classes of information (on 20,30) in the HDTV signal by using a conventional coded modulation scheme and then time-division-multiplexing (in 140) the various coded outputs for transmission.

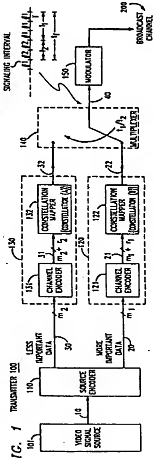

FIG. 1

EP 0 490 552 A2

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Background of the Invention

The present invention relates to the transmission of digital data, particularly the transmission of digital data which represents video signals.

It is generally acknowledged that some form of digital transmission will be required for the next generation of television (TV) technology, conventionally referred to as high definition television, or HDTV. This requirement is due mostly to the fact that much more powerful video compression schemes can be implemented with digital signal processing than with analog signal processing. However, there has been some concern about becoming committed to an all-digital transmission system because of the potential sensitivity of digital transmission to small variations in signal-to-noise ratio, or SNR, at the various receiving locations.

This phenonenon -- sometines referred to as the "threshold effect" -- can be illustrated by considering the case of two television receivers that are respectively located at 50 and 63 miles from a television broadcast station. Since the power of the broadcast signal varies roughly as the inverse square of the distance, it is easily verified that the difference in the amount of signal power received by the television receivers is about 2 dB. Assume, now, that a digital transmission scheme is used and that transmission to the receiver that is 50 miles distant exhibits a bit-error rate of 10^{-6}. If the 2 dB of additional signal loss for the other TV set translated into a 2 dB decrease of the SNR at the input of the receiver, then this receiver will operate with a bit-error rate of about 10^{-4}. With these kinds of bit-error rates, the TV set that is 50 miles away would have a very good reception, whereas reception for the other TV set would probably be very poor. This kind of quick degradation in performance over short distances is generally not considered acceptable by the broadcasting industry. (By comparison, the degradation in performance for presently used analog TV transmission schemes is much more graceful.)

There is thus required a digital transmission scheme adaptable for use in television applications which overcomes this problem. Solutions used in other digital transmission environnents--such as the use of a) regenerative repeaters in cable-based transmission systems or b) fall-back data rates or conditioned telephone lines in voiceband data applications--are clearly inapplicable to the free-space broadcast environment of television.

The co-pending, commonly assigned United States patent application of V. B. Lawrence et al. entitled "Coding for Digital Transmission," serial No. 07/611225, filed on November 07, 1990, discloses a technique for overcoming the shortcomings of standard digital transmission for over-the-air broadcasting of digital TV signals. Specifically, the Lawrence et al. patent application teaches the notion of characterizing the HDTV signal into classes of "less important" and "more important" information which will then use a constellation of non-uniformly spaced signal points. This approach provides unequal error protection, i.e., more error protection for the more important information, and allows a graceful degradation in reception quality at the TV set location because, as the bit-error rate at the receiver begins to increase with increasing distance from the broadcast transmitter, it will be the bits that represent proportionately less of the TV signal information that will be the first to be affected.

Summary of the Invention

Although the Lawrence et al. patent application teaches an advantageous technique for providing unequal error protection to a plurality of classes of information within a signal, we have discovered an alternative, and also advantageous, technique for providing unequal error protection. Specifically, and in accordance with the present invention, unequal error protection is provided for a signal comprised of a plurality of classes of information by a) separately coding each one of the plurality of classes of information using a different coded modulation scheme and b) multiplexing the plurality of coded outputs for transmission.

In accordance with a feature of the invention, uniformly spaced signal points can be used.

In a preferred embodiment of the invention, an HDTV signal is source-encoded to provide a plurality of classes of information. Each class of information is then separately coded by a different, and conventional, coded modulation scheme, e.g., a 4D 8-state trellis code and a uniformly-spaced QAM signal constellation. The coded outputs of the separate coded modulation schemes are then time-division-multiplexed for transmission of the HDTV signal.

Brief Description of the Drawing

In the drawing,

FIG. 1 is a block diagram of an illustrative transmitter embodying the principles of the invention;

FIG. 2 is a block diagram of an illustrative receiver embodying the principles of the invention;

FIGS. 3-4 when taken together, show an illustrative trellis encoder used in the transmitter of FIG. 1;

FIG. 5 shows an embodiment of a multiplexed coded modulation scheme using a 12-QAM signal constellation and a 48-QAM constellation in the transmitter of FIG. 1;

FIG. 6 shows an alternative embodiment of a multiplexed coded modulation scheme using a 12-QAM signal constellation and a 96-QAM

constellation in the transmitter of FIG. 1;

FIG. 7 shows enother alternative embodimant of a multiplexed coded modulation schema using a 16-QAM signal constellation and a 60-QAM constellation in the transmitter of FIG. 1;

FIG. 8 shows a table comparing the nominal coding gains for the three embodiments of FIGS. 5-7; end

FIG. 9 is a block diagram of an Illustrative transmitter embodying the principles of the invention using a concatenated coding technique.

Detalled Descriptlon

Before proceeding with a description of the illustrative embodiment, It should be noted that the various digital signaling concepts described herein--with the exception, of course, of the Inventive concept Itself--are all well known in, for example, the digital radio and voiceband data transmission (modem) arts end thus need not be described in detail herein. These include such concepts as multidimensional signalling using 2N-dimensional chennel symbol constellations, where N Is some integer, trellis coding; fractional coding; scrambling; pessband sheping; equalization; Viterbi, or maximum-likelihood, decoding; etc. These concepts are described in such United States patants as U.S. 3,810,021, issued May 7, 1974 to I. Kalat at al.; U S. 4,015,222, issued Merch 29, 1977 to J. Werner; U.S. 4,170,764, issued October 9, 1979 to J. Salz et al.; U.S. 4,247,940, issued January 27, 1981 to K. H. Mueller et al.; U.S. 4,304,962, issued Decembar 8, 1981 to R. D. Fracassi et al.; U.S. 4,457,004, issued June 26, 1984 to A. Gersho et al.; U.S. 4,489,418, issued December 18, 1984 to J. E. Mazo; U.S. 4,520,490, issued May 28, 1985 to L. -F. Wei; U.S. 4,597, 090, issued June 24, 1986 to G. D. Forney, Jr. and U.S. 4,941,154, issued July 10, 1990 to L. -F. Wei. Additionally, reference can also ba made to "Efficient modulation for band-limited signals", G. D. Forney, Jr. et al., IEEE J. Select. Areas Commun., vol. SAC-2, pp. 632-647, September 1984; "Trellis-coded modulation with multidimensional constellations", L.-F. Wei, IEEE Trans. Inform. Theory, vol. IT-33, pp. 483-501, July 1987; and "Multidimensional constellations - Part I: Introduction, figures of merit, and generalized cross constellations," G. D. Fomey, Jr. & L.-F. Wei, IEEE J. Select. Areas Commun., vol. SAC-7, pp. 877-892, August 1989. All of the above are hereby incorporated by reference.

As previously mentioned, the co-pending, U. S. patent application of V. B. Lawrence et al., serial No. 07/611225, filed on November 7, 1990, discloses a technique for overcoming the shortcomings of standard digital transmission for over-the-air broadcasting of digital TV signals. Specifically, the Lawrence at al. patent application teaches the notion of characterizing the HDTV signal into classes of "less important" and "mora Important" Information which will then use a constellation of non-uniformly speced signal points. This approach provides unequal error protection, I.e., more protection for the more important information, and allows a graceful degradation in reception quality at the TV set location because, as the bit-error rate at the receiver begins to increase with increasing distance from the broadcast transmitter, it will be the bits that represent proportionately less of the TV signal information that will be the first to be affected. However, we have discovered an altemative, also advantageous, technique for providing unequal error protection. Specifically, and In accordance with the present Invention, unequal error protection is provided for a signal comprised of a plurality of classes of information by a) separately coding each one of the plurality of classes of Information using a different coded modulation scheme and b) multiplexing the plurality of coded outputs for transmission. Before proceeding with a description of three Illustrative embodiments of the invention, the Inventive concept Itself will generally be described.

Tuming, In particular, to FIG. 1, Information signal source 101 generates an HDTV analog video signel (HDTV signal) representing picture Information. The HDTV signel is passed on to source encoder 110 which generates a digital signal comprised of a plurality of data elements which are grouped into "classes of information" In which at least one class of Information is more importent, I.e., contains "more important data", than the remainder of the classes of Information which, therefore, contain "less Important data". For example, the more important data represents that Information which, If properly receivad, will form a rough picture, e.g., audio Information, framing information, etc., and the less important data represents that information which comprises the remainder of the HDTV signal. As represented herein, the more Important data Is generated on lead 20 and the less Important data is generated on lead 30. Illustratively, each data element is a data bit, with an average of m_1 (m_2) bits being generated on lead 20 (30) for each signaling Interval assigned by multiplexer 140 to the more (less) important data (see below), each signaling Interval having a duration of T seconds.

As shown in FIG. 1, the more Important data on lead 20 is input to channel encoder 121 of coded modulation circultry 120, and the less important data on lead 30 is Input to channel encoder 131 of coded modulation circuitry 130. Coded modulation circuitry 120 (130) represents a coded modulation scheme and Is comprised of channel encoder 121 (131) and constellation mapper 122 (132). In eccordanca with a principla of the Invention, the coded modulation schemes Implemented by coded modulation circuitry 120 and 130 (described below) are chosen such that the mora Important data is provided more error protection than the less important data, I.e., coded modu-

lation circuitry 120 and 130 are different, with channel encoders 121 and 131, and/or constellation mappings 122 and 132 being different from each other. Channel encoder 121 (131) operates in accordance with known encoding techniques (described below), and the "encoded output" of channel encoder 121 (131) consists of $m_1 + r_1$ ($m_2 + r_2$) data bits, where r_1 (r_2) represents the average number of redundant bits introduced by the encoder 121 (131) in each signaling interval assigned by multiplexer 140 to the more (less) important data. The encoded output of channel encoder 121 (131) is mapped to a signal point from constellation A (B), for each assigned signaling interval, by constellation mapper 122 (132) to provide the "coded output" on leads 22 (32) to multiplexer 140.

Multiplexer 140, illustratively a time-division-multiplexer, is shown as a switch with a design parameter t_1/t_2, i.e., over a time frame $t_f = t_1 + t_2$, multiplexer 140 will switch between coded modulation circuitry 120 and 130. For example, during the time interval t_1 multiplexer 140 will provide the coded output from coded modulation circuitry 120 to modulator 150, and during the time interval t_2 multiplexer 140 will provide the coded output from coded modulation circuitry 130 to modulator 150. (It should be noted that although the simple case of only two classes of information is described herein, the concept can easily be extended to a larger plurality of classes.) Each time interval t_i, for i = 1, 2, is comprised of a number of signaling intervals, T, i.e., $t_1 = N_1T$ and $t_2 = N_2T$, where N_1 (N_2) is the number of signaling intervals in t_1 (t_2). In fact, the design parameter t_1/t_2 denotes the ratio of the numbers of signaling intervals assigned to the more important data and the less important data (i.e., the signaling intervals assigned to channel encoders 121 and 131). For example, for each signaling interval in t_1 (t_2), channel encoder 121 (122) is mapped to a signal point from constellation A (B) so that over the time interval t_1 (t_2) the coded output of coded modulation circuitry 120 (130) will be comprised of N_1 (N_2) signal points. Therefore, and in accordance with the principles of the present invention, by allocating separate time intervals to the more important data and the less important data in a time frame, t_f, the more important data can be separately and differently coded from the less important data. Further, by changing the ratio of t_1/t_2, additional flexibility can be achieved in the design of the separate coding schemes to provide further error protection for the more important data at the expense of the less important data. For example, by increasing the duration of t_1 relative to t_2, the size of the signal constellation used by constellation mapper 122 can be smaller, i.e. the signal points can be spaced further apart, however, this will result in t_2 being smaller, which will require constellation mapper 132 to use a larger constellation of signal points, i.e., the signal points which will be closer together. As a result, since the distance between signal points in a constellation has an effect on the amount of error protection provided by a coded modulation scheme, the error protection of the more important data is enhanced as the expense of the less important data. Coded modulation circuitry 120 and 130, and multiplexer 140 are illustrative of an implementation of a "multiplexed coded modulation scheme". The bandwidth efficiency of the multiplexed coded modulation scheme of FIG. 1 is given by $(m_1t_1 + m_2t_2)/(t_1 + t_2)$ data bits per signaling interval, with the fraction of more important data being $(m_1 t_1)/(m_1t_1 + m_2t_2)$ of the total. The coded outputs from the multiplexed coded modulation scheme are provided to modulator 150, which is representative of conventional television broadcasting circuitry, for transmission of the broadcast HDTV signal on broadcast channel 200.

The broadcast HDTV signal is received from broadcast channel 200 by receiver 300 which is shown in FIG. 2. The broadcast HDTV signal is received by demodulator 350 which is representative of conventional reception and demodulation circuitry, e.g., the antenna, demodulation, analog-to-digital conversion, etc. Demodulator 350 provides a time-multiplexed digital signal representing the received coded outputs on lead 90 which is processed by demultiplexer 340 to provide the separate received coded outputs. The received coded output representing the more important data is provided to channel decoder 331 and the received coded output representing the less important data is provided to channel decoder 332. Channel decoder 331 (332) decodes the received coded output representing the more important (less important) data to provide the more important (less important) data, i.e., class of information, to source decoder 310. Source decoder 310 provides the inverse function of source encoder 110 of transmitter 100 to provide the received HDTV signal to CRT display 301.

Having described the general inventive concept above, various illustrative embodiments of a multiplexed coded modulation scheme will now be described. Although any coded modulation scheme can be implemented in coded modulation circuitry 120 and 130, the present invention advantageously allows the use of simple channel encoders and constellations of uniformly spaced signal points. For the remainder of the discussion, it is assumed that channel encoders 121 and 131 are implemented using a simple 4D 8-state trellis encoder as shown in FIGS. 3-4 (in FIG. 3, the boxes labeled "T" are T-second delay elements, the circles labeled "+" are exclusive-or gates, and the bit-converter operates in accordance with FIG. 4). Further, it will be assumed that signal constellations 122 and 132 are representative of uniformly-spaced QAM constellations and, although differing in size (i.e., numbers of signal points), have the same average power (average energy per signal point).

FIGS. 5 - 7 illustrate a variety of embodiments of an illustrative multiplexed coded modulation scheme for different a) values of m_1 and m_2, b) QAM signal constellations, and c) t_1/t_2 multiplexer ratios. FIG. 8 lists various characteristics of these embodiments. The bandwidth efficiency each of these embodiments is four data bits per signaling interval, with the percentage of more important data varying from 37.5% to 62.5% of the total. (It should be noted that these embodiments are for comparison purposes only, e.g., other bandwidth efficiencies can be used, different signal constellations can be used (with different sizes), etc.) For example, applying the above mentioned bandwidth efficiency formula to the embodiment shown in FIG. 5, i.e., ($m_1 = 3$, $m_2 = 5$), and ($t_1 = t_2 = T$) yields four data bits per signaling interval:

$$\frac{(m_1 t_1 + m_2 t_2)}{t_1 + t_2} = \frac{3T + 5T}{T + T} = 4$$

In each embodiment, the sizes of the signal constellations and the nominal coding gains for the more important and less important data are determined based on the above assumption that the simple 4D 8-state trellis code of FIGS. 3-4 is used in both channel encoders 121 and 131.

It should be observed in FIG. 3 that two input bits are coded every two signal intervals to provide three encoded bits (i.e., the delay element of the 4D 8-state trellis code is 2T signaling intervals). The three encoded bits, together with an uncoded input bit, are then converted into two pairs of output bits through the bit converter of FIG. 4. Each pair of output bits is next used to identify, in the first or second signaling intervals, one out of four 2D subsets of signal points, as shown by the example of constellation (A) in FIG. 5, where each subset identified by a two bit pattern consists of these signal points. The four 2D subsets are obtained by partitioning the corresponding constellation so that the distance between the signal points in each subset is greater than that between the signal points of the overall constellations, as in the conventional coded modulation. Any number of input bits in excess of three will remain uncoded and be used to select a 2D signal point from each of the two identified 2D subsets (some processing on the uncoded bits may be needed in order to simplify the selection process, e.g., see United States patent 4,941, 154, issued July 10, 1990 to L. -F. Wei, and "Multidimensional constellations - Part I: Introduction, figures of merit, and generalized cross constellations," G. D. Forney, Jr. & L.-F. Wei, IEEE J. Select. Areas Commun., vol. SAC-7, pp. 877-892, August 1989).

In each embodiment the real coding gain is expected to be less than its corresponding nominal coding gain, which is due to the large error coefficient associated with the Minimum Squared Euclidean Distance (MSED) of the 4D 8-state trellis code. The Peak-to-Average Power Ratio (PAR) of the three embodiments are determined by the larger constellations used for the less important data, which are all slightly bigger than two.

It may also be noted that additional coded modulation schemes can be implemented within a multiplexed coded modulation scheme to protect against other forms of noise that may be present in a communications system. For example, the conventional coded modulation schemes used in FIGS. 5-7 are not effective against impulse noise, so a well-known Reed-Solomon code which is effective against impulse noise can be used in conjunction with a trellis code to form a concatenated code. A block diagram of an illustrative embodiment using a concatenated code is shown in FIG. 9. In FIG. 9, the more (less) important data is first separately encoded by first channel encoder 115 (116) which uses a well-known Reed-Solomon code (i.e., additional redundant bits are added to m_1 (m_2)), and then further encoded by second channel encoder 121 (131) using the trellis code described above (it should be noted that channel encoder 121 (131) and constellation mapper 122 (132) have to be modified accordingly to handle the additional redundant bits introduced by first channel encoder 115 (116)).

The foregoing merely illustrates the principles of the invention. For example, although the invention is illustrated herein as being implemented with discrete functional building blocks, e.g., source decoders, channel encoders, etc., the functions of any one or more of those building blocks can be carried out using one or more appropriate programmed processors, digital signal processing (DSP) chips, etc. In addition, the invention could be implemented such that some of the discrete functional blocks were shared in time, e.g., physically using only one channel encoder that is switched between two signal constellations. Also, the coded modulation scheme for each class of information can be enhanced using interleaving techniques, or more complex coded modulation schemes, to protect against other forms of noise, e.g., to protect against "colored" noise. Further, other multiplexing techniques may be used in place of time-division-multiplexing.

It will thus be appreciated that those skilled in the art will be able to devise numerous and various alternative arrangements which, although not explicitly shown or described herein, embody the principles of the invention and are within its spirit and scope.

Claims

1. A method for processing an information signal, the information signal being comprised of a plurality of classes of information, the method characterized by the steps of:

separately coding each one of the plurality

of classes of information using a separate coded moduletion scheme to provide a coded output, such that one of the plurality of classes of information has more error protection than the remaining ones of the plurality of classes of information; and

multiplexing the plurality of coded outputs for transmission.

2. The method of claim 1 characterized in that the multiplexing step is time-division-multiplexing.

3. The method of claim 2 characterized in that the multiplexing step includes the step of assigning each one of the plurelity of coded outputs to a time interval in a time frame, the time frame being greater than or equal to the sum of the plurality of assigned time intervals.

4. The method of claim 1 characterized in that the separetely coding step includes the steps of:

encoding each one of the plurality of classes of information to provide an encoded output; and

mapping each one of the plurality of encoded outputs to a signal point of e signal constellation to provide the coded output.

5. The method of claim 4 characterized in that at least one of the plurality of encoded outputs is mapped to a signal constellation of different size than the remaining ones of the plurality of encoded outputs.

6. A method for providing unequal error protection for an information signai, the information signal being comprised of a plurality of classes of information, the method characterized by the steps of:

assigning each one of the plurality of classes of information to a coded moduiation scheme such that at least one of the plurality of classes of information is assigned to a different coded modulation scheme than the remaining ones of the plurality of classes of information;

assigning a time interval to each one of the plurality of coded moduletion schemes such that at least one of the plurality of coded modulation schemes is assigned to a different time interval than the remaining ones of the plurality of coded modulation schemes; and

separately coding each one of the plurality of classes of information using the assigned coded modulation scheme in the essigned time interval to provide a coded output for transmission such that at least one of the plurality of classes of information has more error protection than the remaining ones of the piurality of classes of information.

7. Apparatus for processing an information signal, the information signel being comprised of a plurality of classes of information, the apparatus being characterized by:

source encoding means responsive to the information signal for providing the plurality of classes of information;

coding means responsive to the plurality of classes of information for separately coding each one of the plurality of classes of information using a separate coded modulation scheme to provide a coded output for each one of the plurality of classes of information such that at least one of the plurality of classes of information has more error protection then the remaining ones of the plurality of classes of information; and

means for multiplexing the plurality of coded outputs for transmission.

8. The apparetus of claim 7 characterized in that the means for multiplexing operates in accordance with time-division-multiplexing.

9. The apparatus of claim 7 cheracterized in that the means for multiplexing assigns each one of the plurality of coded outputs to a time interval in a time freme, the time frame being greater than or equal to the sum of the plurality of essigned time intervals.

10. The apparatus of claim 7 characterized in that the coding means is further comprised of:

means for channel encoding each one of the plurality of clesses of information to provide an encoded output; and

means for mapping each one of the encoded outputs to a signal point of a signal constellation to provide the coded output.

11. The apparatus of claim 10 characterized by at least one of the plurality of encoded outputs is mapped to a signal constellation of different size than the remaining ones of the plurality of encoded outputs.

12. Apparatus for decoding a received signal, the received signal being characterized by a plurality of coded outputs, each one of the plurality of coded outputs representing a class of information and where at least one class of information is provided more error protection than the remaining ones of the plurality of classes of information, the apparatus being comprised of:

means for demultiplexing the received signal to provide the plurality of coded outputs;

means for decoding each one of the plurality of coded outputs using a separate decoding scheme to provide each one of the clas-

ses of information; and

means for source decoding the plurality of classes of information to provide an information signal.

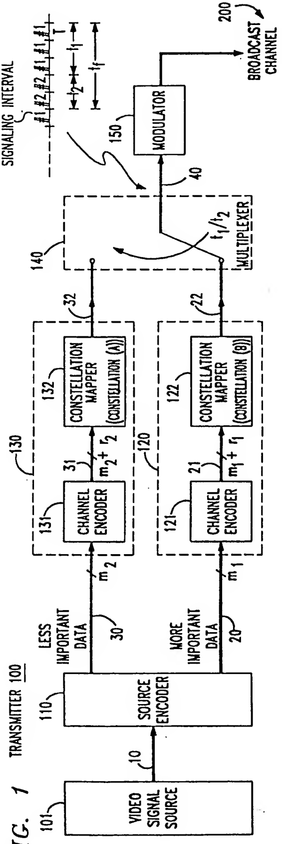

TRANSMITTER 100
101
VIDEO SIGNAL SOURCE
10
110
SOURCE ENCODER
LESS IMPORTANT DATA
30
MORE IMPORTANT DATA
20
m2
m1
130
131
CHANNEL ENCODER
31
m2+ r2
132
CONSTELLATION MAPPER
(CONSTELLATION (A))
32
120
121
CHANNEL ENCODER
21
m1+ r1
122
CONSTELLATION MAPPER
(CONSTELLATION (B))
22
140
MULTIPLEXER
t1/t2
40
150
MODULATOR
BROADCAST CHANNEL
200
SIGNALING INTERVAL
#1 #2 #2 #2 #1 #1 #1 #1
T
t2
t1
tf

FIG. 1

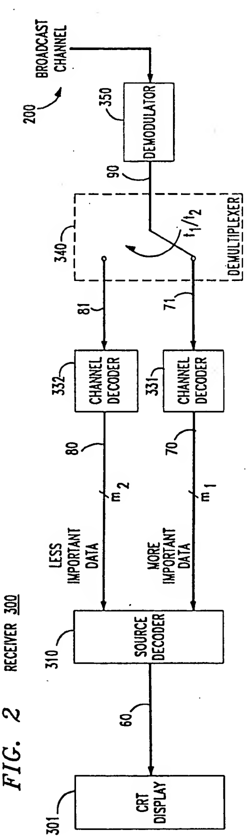

BROADCAST CHANNEL
200
350
DEMODULATOR
90
340
DEMULTIPLEXER
t1/t2
81
71
332
CHANNEL DECODER
331
CHANNEL DECODER
80
70
m2
m1
LESS IMPORTANT DATA
MORE IMPORTANT DATA
RECEIVER 300
310
SOURCE DECODER
60
301
CRT DISPLAY

FIG. 2

FIG. 3

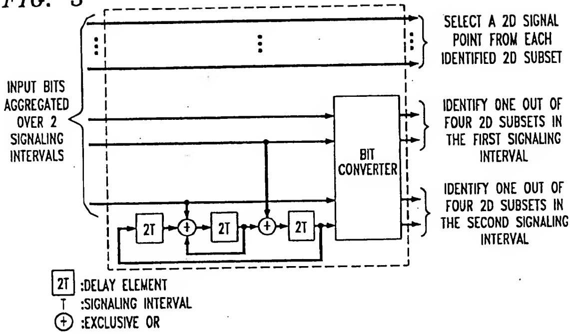


FIG. 4

BIT CONVERTER IN FIG. 3	
INPUT BIT PATTERN*	OUTPUT BIT PATTERN*
0 0 0 0	0 0 0 0
0 0 0 1	0 0 0 1
0 0 1 0	0 0 1 1
0 0 1 1	0 0 1 0
0 1 0 0	0 1 0 1
0 1 0 1	0 1 1 1
0 1 1 0	0 1 1 0
0 1 1 1	0 1 0 0
1 0 0 0	1 1 1 1
1 0 0 1	1 1 1 0
1 0 1 0	1 1 0 0
1 0 1 1	1 1 0 1
1 1 0 0	1 0 1 0
1 1 0 1	1 0 0 0
1 1 1 0	1 0 0 1
1 1 1 1	1 0 1 1

*READING FROM TOP TO BOTTOM IN FIG. 3

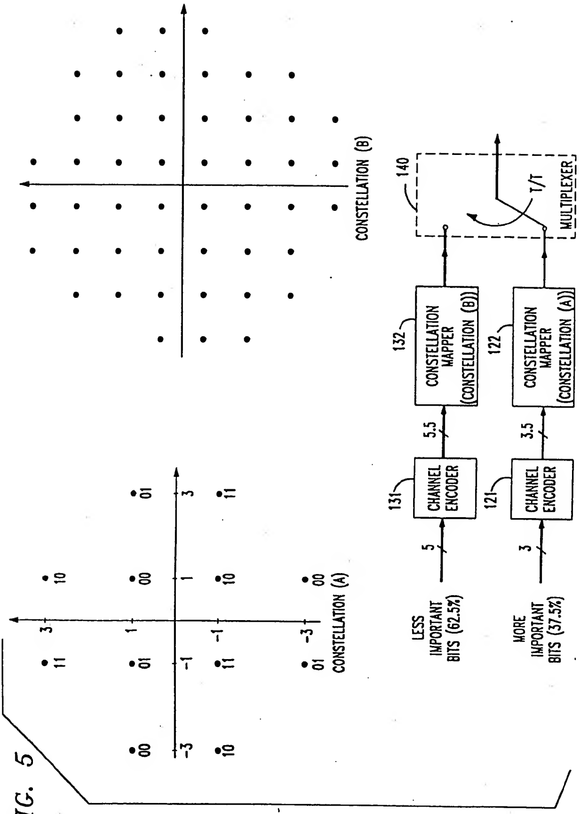

CONSTELLATION (A)
CONSTELLATION (B)
LESS IMPORTANT BITS (62.5%)
MORE IMPORTANT BITS (37.5%)
5
3
5.5
3.5
131
121
CHANNEL ENCODER
CHANNEL ENCODER
132
122
CONSTELLATION MAPPER (CONSTELLATION (B))
CONSTELLATION MAPPER (CONSTELLATION (A))
140
T/T
MULTIPLEXER

FIG. 5

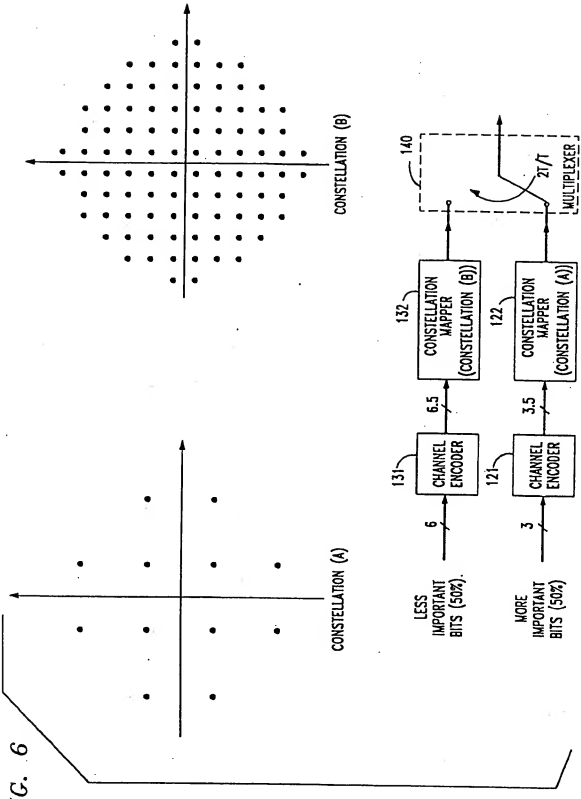

CONSTELLATION (A)
CONSTELLATION (B)
LESS IMPORTANT BITS (50%)
MORE IMPORTANT BITS (50%)
6
3
131
121
CHANNEL ENCODER
CHANNEL ENCODER
6.5
3.5
132
122
CONSTELLATION MAPPER (CONSTELLATION (B))
CONSTELLATION MAPPER (CONSTELLATION (A))
140
2T/T
MULTIPLEXER

FIG. 6

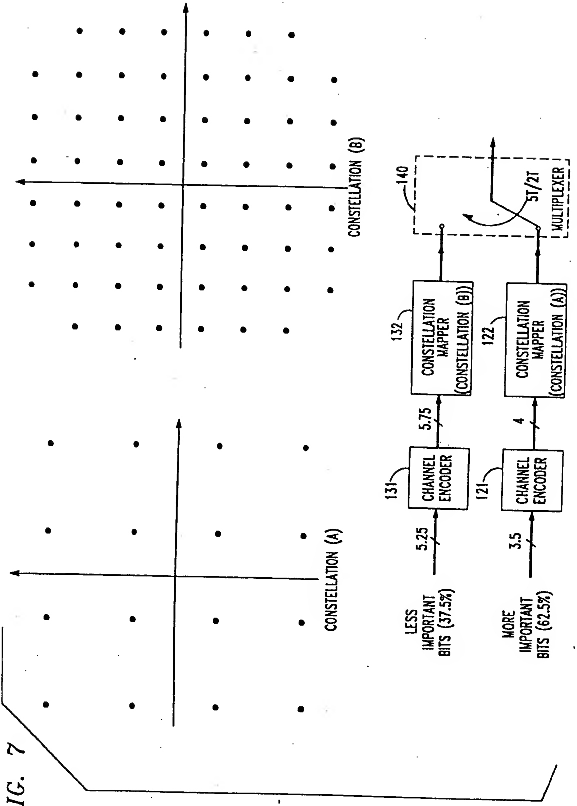

CONSTELLATION (A)
CONSTELLATION (B)
LESS IMPORTANT BITS (37.5%)
MORE IMPORTANT BITS (62.5%)
5.25
3.5
131
121
CHANNEL ENCODER
CHANNEL ENCODER
5.75
4
132
122
CONSTELLATION MAPPER (CONSTELLATION (B))
CONSTELLATION MAPPER (CONSTELLATION (A))
140
5T/2T
MULTIPLEXER

FIG. 7

FIG. 8

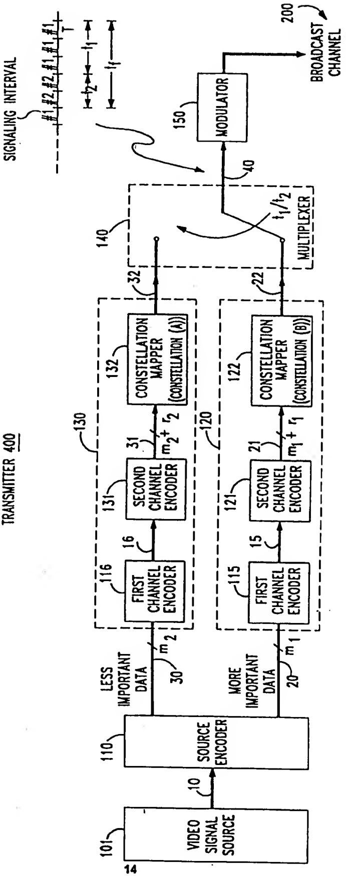
COMPARISONS OF MULTIPLEXED CODED MODULTATION

SCHEME	PERCENTAGE OF IMPORTANT DATA (%)	MORE IMPORTANT DATA			LESS IMPORTANT DATA			t_1/t_2 OF MULTIPLEXER	PEAK-TO-AVERAGE POWER RATIO
		m_1	CONSTELLATION (A)	NOMINAL CODING GAIN (dB)**	m_2	CONSTELLATION (B)	NOMINAL CODING GAIN (dB)**		
1	37.5	3	12-QAM	7.6	5	48-QAM	1.5	T/T	2.07
3	50	3	12-QAM	7.6	6	96-QAM	-1.5	2T/T	2.17
4	62.5	3.5	16-QAM	6.0	5.25	60-QAM	0.8	5T/2T	2.24

** RELATIVE TO UNCODED 16-QAM

FIG. 9

TRANSMITTER 400



P/0568-07

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(54) Communication system using trellis coded QAM.

(57) Coded modulation schemes based on codes for QPSK modulation are directly incorporated into QAM based modulation systems, forming trellis coded QAM, to provide a practical coding structure that is both efficient in bandwidth and data reliability. Concatenated coding with QPSK based trellis coding and symbol error correcting coding is used. In an encoder (Fig. 2), an N-bit QAM constellation pattern (80) is divided into four subsets, each including N/4 symbol points of the constellation pattern. A two-bit QPSK codeword (92) is assigned to each of the four subsets (82, 84, 86, 88). A symbol to be transmitted is first encoded using an outer error correcting encoding algorithm, such as a Reed-Solomon code (12). Part of the symbol is then encoded (48) with an inner code that comprises a rate 1/2 trellis encoding algorithm to provide a QPSK codeword, which is mapped (50) with the remaining bits of the symbol to provide a modulation function, wherein the remaining bits (94) correlate the symbol with one of the symbol points included in the subset defined by the QPSK codeword. At a receiver (Fig. 3), the recovered modulation function is pruned (62) to provide a set of metrics (66) corresponding to the subsets and to provide a plurality of conditional determinations of the constellation point identified by the remaining bits. The metrics are used in a rate 1/2 trellis decoder (68) to recover a first bit that is encoded using a rate 1/2 encoding algorithm to recreate the QPSK codeword. One of a plurality of the conditional determinations is selected in response to the recreated codeword. The selected conditional determination is combined with the recovered first bit to provide a decoded output that is further decoded using a symbol error correcting algorithm such as a Reed-Solomon code (36).

EP 0 525 641 A2

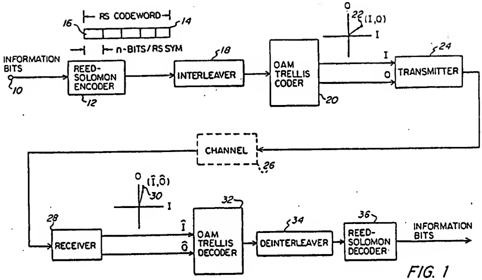

RS CODEWORD
16
14
n-BITS/RS SYM
INFORMATION BITS
10
REED-SOLOMON ENCODER
12
18
INTERLEAVER
QAM TRELLIS CODER
20
Q
22
(I,Q)
I
I
Q
24
TRANSMITTER
CHANNEL
26
Q
(Î,Q̂)
30
I
28
RECEIVER
Î
Q̂
32
QAM TRELLIS DECODER
34
DEINTERLEAVER
36
REED-SOLOMON DECODER
INFORMATION BITS

FIG. 1

BACKGROUND OF THE INVENTION

The present invention relates to trellis coded quadrature amplitude modulation (QAM) and more particularly to a practical method for coding QAM transmission.

Digital data, for example digitized video for use in broadcasting high definition television (HDTV) signals, can be transmitted over terrestrial VHF or UHF analog channels for communication to end users. Analog channels deliver corrupted and transformed versions of their input waveforms. Corruption of the waveform, usually statistical, may be additive and/or multiplicative, because of possible background thermal noise, impulse noise, and fades. Transformations performed by the channel are frequency translation, nonlinear or harmonic distortion and time dispersion.

In order to communicate digital data via an analog channel, the data is modulated using, for example, a form of pulse amplitude modulation (PAM). Typically, quadrature amplitude modulation (QAM) is used to increase the amount of data that can be transmitted within an available channel bandwidth. QAM is a form of PAM in which a plurality of bits of information are transmitted together in a pattern referred to as a "constellation" that can contain, for example, sixteen or thirty-two points.

In pulse amplitude modulation, each signal is a pulse whose amplitude level is determined by a transmitted symbol. In 16-bit QAM, symbol amplitudes of -3, -1, 1 and 3 in each quadrature channel are typically used. Bandwidth efficiency in digital communication systems is defined as the number of transmitted bits per second per unit of bandwidth, i.e., the ratio of the data rate to the bandwidth. Modulation systems with high bandwidth efficiency are employed in applications that have high data rates and small bandwidth occupancy requirements. QAM provides bandwidth efficient modulation.

On the other hand, modulation schemes such as quadrature phase shift keying (QPSK), commonly found in satellite transmission systems, are well established and understood. In QPSK, a more simple constellation pattern than that provided in QAM results. In particular, QPSK systems use a constellation pattern having only four symbols that are typically positioned 90 degrees apart from each other in phase, but have the same amplitude. Thus, the four symbols are equally spaced about a circle.

QPSK modulation is suitable for power limited systems where bandwidth limitations are not a major concern. QAM modulation, on the other hand, is advantageous in bandwidth limited systems, where power requirements do not present a major problem. Therefore, QPSK has been the system of choice in satellite communication systems, whereas QAM is preferred in terrestrial and cable systems. As a consequence of the popularity of QPSK, integrated circuits that realize trellis coded QPSK modulation are readily available and easily obtained.

Trellis coded modulation (TCM) has evolved as a combined coding and modulation technique for digital transmission over band limited channels. It allows the achievement of significent coding gains over conventional uncoded multilevel modulation, such as QAM, without compromising bandwidth efficiency. TCM schemes utilize redundant nonbinary modulation in combination with a finite-state encoder which governs the selection of modulation signals to generate coded signal sequences. In the receiver, the noisy signals are decoded by a soft-decision maximum likelihood sequence decoder. Such schemes can improve the robustness of digital transmission against additive noise by 3-6 dB or more, compared to conventional uncoded modulation. These gains are obtained without bandwidth expansion or reduction of the effective information rate as required by other known error correction schemes. The term "trellis" is used because these schemes can be described by a state-transition (trellis) diagram similar to the trellis diagrams of binary convolutional codes. The difference is that TCM extends the principles of convolutional coding to nonbinary modulation with signal sets of arbitrary size.

The availability of components for implementing trellis coded QPSK modulation is a significant advantage in designing low cost communication systems for applications, such as satellite communications, wherein QPSK techniques excel. However, such components have not been of assistance in implementing other coded transmission systems, such as those in which QAM is preferred.

For applications that are both power limited and band limited, and require low cost components (particularly low cost data decoders), conventional QAM systems have not been feasible due to the complexity and relatively high cost of the required encoder and decoder circuits. In fact, it is typical to implement QAM trellis encoders and decoders in expensive custom integrated circuit chips.

One power limited and band limited application in which a low cost solution is necessary for communicating digital data is the digital communication of compressed high definition television signals. Systems for transmitting compressed HDTV signals have data rate requirements on the order of 15-20 megabits per second (Mbps), bandwidth occupancy requirements on the order of 5-6 MHz (the bandwidth of a conventional National Television System Committee (NTSC) television channel), and very high data reliability requirements (i.e., a very small bit error rate). The data rate requirement arises from the need to

provide a high quality compressed television picture. The bandwidth constraint is a consequence of the U.S. Federal Communications Commission requirement that HDTV signals occupy existing 6 MHz television channels, and must coexist with the current broadcast NTSC signals. This combination of data rate and bandwidth occupancy requires a modulation system that has high bandwidth efficiency. Indeed, the ratio of data rate to bandwidth must be on the order of 3 or 4. This means that modulation systems such as QPSK, having a bandwidth efficiency without coding of two, are unsuitable. A more bandwidth efficient modulation, such as QAM is required. However, as noted above, QAM systems have been too expensive to implement for high volume consumer applications.

The requirement for a very high data reliability in the HDTV application results from the fact that highly compressed source material (i.e., the compressed video) is intolerant of channel errors. The natural redundancy of the signal has been removed in order to obtain a concise description of the intrinsic value of the data. For example, for a system to transmit at 15 Mbps for a twenty-four hour period, with less than one bit error, requires the bit error rate (BER) of the system to be less than one error in 10^{12} transmitted bits.

Data reliability requirements are often met in practice via the use of a concatenated coding approach, which is a divide and concur approach to problem solving. In such a coding framework, two codes are employed. An "inner" modulation code cleans up the channel and delivers a modest symbol error rate to an "outer" decoder. The inner code is usually a coded modulation that can be effectively decoded using "soft decisions" (i.e., finely quantized channel data). A known approach is to use a convolutional or trellis code as the inner code with some form of the "Viterbi algorithm" as a trellis decoder. The outer code is most often a t-error-correcting, "Reed-Solomon" code. Such Reed-Solomon coding systems, that operate in the data rate range required for communicating HDTV data, are widely available and have been implemented in the integrated circuits of several vendors. The outer decoder removes the vast majority of symbol errors that have eluded the inner decoder in such a way that the final output error rate is extremely small.

A more detailed explanation of concatenated coding schemes can be found in G. C. Clark, Jr. and J. B. Cain, "Error-Correction Coding for Digital Communications", Plenum Press, New York, 1981; and S. Lin and D. J. Costello, Jr., "Error Control Coding: Fundamentals and Applications", Prentice-Hall, Englewood Cliffs, New Jersey, 1983. Trellis coding is discussed extensively in G. Ungerboeck, "Channel Coding with Multilevel/Phase Signals", IEEE Transactions on Information Theory, Vol. IT-28, No. 1, pp. 55-67, January 1982; G. Ungerboeck, "Trellis-Coded Modulation with Redundant Signal Sets -- Part I: Introduction, -- Part II: State of the Art", IEEE Communications Magazine, Vol. 25, No. 2, pp. 5-21, February 1987; and A. R. Caulderbank and N. J. A. Sloane, "New Trellis Codes Based on Lattices and Cosets", IEEE Transactions on Information Theory, Vol. IT-33, No. 2, pp. 177-195, March 1987. The Viterbi algorithm is explained in G. D. Forney, Jr., "The Viterbi Algorithm", Proceedings of the IEEE, Vol. 61, No. 3, March 1973. Reed-Solomon coding systems are discussed in the Clark, Jr. et al and Lin et al articles cited above.

The error rate performance at the output of the inner, modulation code in concatenated coded systems is highly dependent on signal-to-noise ratio (SNR). Some codes perform better, providing a lower error rate at a low SNR while others perform better at a high SNR. This means that the optimization of the modulation code for concatenated and nonconcatenated coding systems can lead to different solutions, depending on the specified SNR range.

It would be advantageous to provide a data modulation system with high bandwidth efficiency and low power requirements. Such a system should provide a high data rate, with minimal bandwidth occupancy, and very high data reliability. The complexity of a receiver for use with such a system should be minimized, to provide low cost in volume production. Optimally, the system should be able to be implemented using readily available components with as little customization as possible.

The present invention provides a modulation system having the aforementioned advantages. In particular, the method and apparatus of the present invention expand a trellis coded QPSK system to a trellis coded QAM system, without sacrificing data reliability.

SUMMARY OF THE INVENTION

In accordance with the present invention, a method is provided for communicating digital data using QAM transmission. An n-bit QAM constellation pattern is divided into four subsets. Each subset includes N/4 symbol points of the constellation pattern. A different two-bit codeword is assigned to each of the four subsets. A symbol to be transmitted is encoded by processing a first bit of the symbol with a rate 1/2 binary convolutional encoding algorithm to provide the two-bit codeword assigned to the subset in which the symbol resides in the constellation pattern. The two-bit codeword is mapped with the remaining bits of the symbol to provide a modulation function. The remaining bits correlate the symbol with one of the N/4

symbol points included in the subset defined by the codeword. A carrier is modulated with the modulation function for transmission on a communication channel.

In an illustrated embodiment, the two-bit codeword forms the least significant bits of the modulation function and defines the columns of a matrix of coordinates of the constellation pattern. The remaining bits form the most significant bits of the modulation function and determine the size of the constellation pattern. In a concatenated approach, information bits are first encoded into symbols using, for example, a t-symbol error correcting code, such as a Reed-Solomon code. These encoded symbols are then passed to a trellis encoder which produces the desired modulation for a carrier.

After the modulation function is transmitted, it is recovered at a receiver. The recovered modulation function is pruned to provide a set of metrics corresponding to the subsets and to provide a plurality of bytes representing different conditional determinations of a signal point identified by the remaining bits. The metrics are used in an algorithm (such as the Viterbi algorithm) for decoding a rate 1/2 binary convolutional code to recover the first bit. The recovered first bit is encoded using a rate 1/2 binary convolutional encoding algorithm to recreate the codeword. One of the conditional determination bytes is selected in response to the recreated codeword. The selected byte is then combined with the recovered first bit to provide a decoded output.

The present invention also provides apparatus for encoding digital data for QAM transmission. The encoder includes means for parsing a symbol to be transmitted into a first bit and at least one remaining bit. Means are provided for encoding the first bit with a rate 1/2 binary convolutional encoding algorithm to provide a two-bit codeword that defines one of four subsets of an N-bit QAM constellation pattern, each subset including N/4 symbol points of the constellation pattern. The codeword is mapped with the remaining bits to provide a modulation function. The remaining bits correlate the symbol with one of the N/4 symbol points included in the subset defined by the codeword. Means are provided for modulating a carrier with the modulation function for transmission on a communication channel. An outer encoder can be provided for encoding information bits using an error correcting algorithm to provide the symbol that is parsed by the parsing means.

In an illustrated embodiment, the codeword forms the least significant bits of the modulation function and defines the columns of a matrix of coordinates of said constellation pattern. The remaining bits form the most significant bits of the modulation function and determine the size of the constellation pattern. The encoding means can use a trellis coding algorithm.

Decoding apparatus is also provided in accordance with the invention. A receiver demodulates a received carrier to recover an N-bit QAM modulation function in which a two-bit codeword identifies one of a plurality of QAM constellation subsets and the remaining (N-2) bit portion represents a signal point within said one subset. Means are provided for pruning the recovered modulation function to provide a set of metrics corresponding to said subsets and to provide a plurality of (N-2) bit subgroups representing a plurality of conditional determinations of the signal point identified by the (N-2) bit portion. The metrics are used in an algorithm for decoding a rate 1/2 binary convolutional code to recover a first bit. The recovered first bit is encoded using a rate 1/2 binary convolutional encoding algorithm to recreate the codeword. Means are provided for selecting one of the plurality of (N-2) bit subgroups in response to the recreated codeword. The selected subgroup is combined with the recovered first bit to provide a decoded output.

In an illustrated embodiment, the codeword comprises the least significant bits in the modulation function and defines the columns of a matrix of constellation coordinates, with the selected subgroup forming the most significant bits and defining a row of the matrix. The pruning means quantize the recovered N-bit modulation function for each column of a matrix of constellation coordinates and the conditional determinations comprise a best choice for each of the columns with the set of metrics identifying the quality of each choice. The metrics are used in conjunction with a decoder that uses a soft-decision algorithm for decoding convolutional codes.

A concatenated decoder is also provided. In the concatenated embodiment, an outer decoder is provided for decoding the output using a symbol error correcting algorithm. In an illustrated embodiment, the inner decoding algorithm used in the concatenated decoder comprises the Viterbi algorithm. The outer, symbol error correcting algorithm can comprise a Reed-Solomon code. The carrier signal received by the receiver can comprise a high definition television carrier signal.

BRIEF DESCRIPTION OF THE DRAWINGS

Figure 1 is a block diagram of a QAM transmission system employing concatenated coding;

Figure 2 is a block diagram of a trellis encoder in accordance with the present invention;

Figure 3 is a block diagram of a trellis decoder in accordance with the present invention;

Figure 4 is an illustration of a QAM constellation pattern divided into subsets in accordance with the present invention;
Figure 5 is a diagram defining the labeling of subsets in the constellation pattern of Figure 4;
Figure 6 is a diagram illustrating the labeling of constellation points in the constellation pattern of Figure 4; and
Figure 7 is a graph illustrating the performance of a concatenated coding scheme in accordance with the present invention as compared to a prior art coded QAM scheme.

DETAILED DESCRIPTION OF THE INVENTION

Figure 1 illustrates a concatenated coding system for communicating QAM data. Digital information to be transmitted is input to a symbol error correcting coder 12, such as a Reed-Solomon encoder, via an input terminal 10. Encoder 12 converts the information into a codeword 14, comprising a plurality of successive n-bit symbols 16. While an outer convolutional code could be used for encoder 12, the bursty nature of the errors in a transmission system, the fact that only hard quantized data is available, and the desirability of a high rate code make a Reed-Solomon code, whose symbols are formed from n-bit segments of the binary stream, a good choice for the outer code. Since the performance of a Reed-Solomon code only depends on the number of symbol errors in the block, such a code is undisturbed by burst errors within an n-bit symbol. However, the concatenated system performance is severely degraded by long bursts of symbol errors. Therefore, an interleaver 18 is provided at the output of Reed-Solomon encoder 12, to interleave the symbols (as opposed to individual bits) between coding operations. The intent of the interleaving is to break up the bursts of symbol errors.

The interleaved symbols are input to a QAM trellis coder 20. In accordance with the present invention, coder 20 incorporates a QPSK code into a trellis coded QAM modulation system, as described in greater detail below.

The output of coder 20 comprises symbols representative of coordinates in the real (I) and imaginary (Q) planes of a QAM constellation pattern. One such constellation point 22 is symbolically illustrated in Figure 1. The symbols are transmitted by a conventional transmitter 24 via a communication channel 26. The communication channel introduces various distortions and delays that corrupt the signal before it is received by a receiver 28. As a result, the coordinate values embodied in the received symbols will not correlate exactly with the transmitted coordinate values, such that a received point 30 will end up on the constellation pattern in a different location than the actual transmitted point 22. In order to determine the correct location for the received point, and thereby obtain the data as actually transmitted, the received data ($\tilde{I}$, $\tilde{Q}$) is input to a QAM trellis decoder 32 that uses a soft-decision convolutional decoding algorithm to recover the transmitted information. A decoder in accordance with the present invention is described in greater detail below.

The decoded output from decoder 32 is input to a deinterleaver 34 that reverses the effects of interleaver 18 discussed above. The deinterleaved data is input to a Reed-Solomon decoder 36 for recovery of the original information bits.

In the present invention, a QPSK code is incorporated into the trellis coded QAM modulation system to provide a high data rate, bandwidth efficient system with a moderate bit error rate in low SNR regions of operation. In order to achieve this result, the codewords of the QPSK code and the "uncoded" bits which together define a symbol are uniquely assigned to a QAM constellation. In addition, the received signal is decoded by a combination of a soft-decision decoder with techniques for deciding which constellation points the "uncoded" bits refer to.

Figure 2 illustrates an encoder in accordance with the present invention. Data bits (e.g., from interleaver 18 - Figure 1) are input to a conventional parsing circuit 42 via an input terminal 40. An N-1 bit symbol to be transmitted is parsed into a first bit that is output on line 46 to a convolutional encoder 48. The remaining N-2 "uncoded" bits are output on line 44 to a 2^N-QAM mapper 50. Convolutional encoder 48 employs a rate 1/2, 64-state convolutional code, in which the generators are 171 and 133 in octal. The two bits output from encoder 48 and the N-2 uncoded bits (N bits total) are presented to the 2^N-QAM mapper for use as labels to map the N-bit symbol to a specific constellation point on a QAM constellation. The two "coded" bits output from convolutional encoder 48 are actually QPSK codewords, and are used to select a constellation subset. The uncoded bits are used to select a specific signal point within the constellation subset from the QAM constellation.

For purposes of QAM transmission (encoding), the codewords of the QPSk code and the remaining uncoded bits must be assigned to the QAM constellation. For this purpose, one must describe a labeling of QAM constellation points by a modulation function, $MOD(m) \epsilon R^2$.

$MOD:\{0, 1\}^N \rightarrow R^2.$

The mapping described below has the following desirable features: (1) the consequences of the 90° phase ambiguity of QAM is imposed on the QPSK codewords while the uncoded bits are invariant to the ambiguity (i.e., the 90° phase ambiguity can be dealt with in the same manner es the QPSK system) and (2) the most significant digits control the constellation size (i.e, a nested scheme for 16/32/64 - QAM).

Consider the labeling described by the following matrix, for 16-QAM ($m_5 = m_4 = 0$) (and QPSK, $m_5 = m_4 = M_3 = m_2 = 0$):

$$\begin{array}{l} MOD(m_5 m_4 m_3 m_2 m_1 m_0) \qquad\qquad m_1 m_0 \\ \begin{array}{c} m_5 m_4 m_3 m_2 \\ 0\,0\,0\,0 \\ 0\,0\,0\,1 \\ 0\,0\,1\,1 \\ 0\,0\,1\,0 \end{array} \begin{array}{c} \begin{array}{cccc} 0\,0 & 0\,1 & 1\,1 & 1\,0 \end{array} \\ \begin{pmatrix} +1,+1 & -1,+1 & -1,-1 & +1,-1 \\ +1,-3 & +3,+1 & -1,+3 & -3,-1 \\ -3,-3 & +3,-3 & +3,+3 & -3,+3 \\ -3,+1 & -1,-3 & +3,-1 & +1,+3 \end{pmatrix} \end{array} \end{array},$$

for 32-QAM ($m_5 = 0$) add:

$$\begin{array}{l} MOD(m_5 m_4 m_3 m_2 m_1 m_0) \qquad\qquad m_1 m_0 \\ \begin{array}{c} m_5 m_4 m_3 m_2 \\ 0\,1\,0\,0 \\ 0\,1\,0\,1 \\ 0\,1\,1\,1 \\ 0\,1\,1\,0 \end{array} \begin{array}{c} \begin{array}{cccc} 0\,0 & 0\,1 & 1\,1 & 1\,0 \end{array} \\ \begin{pmatrix} +5,-3 & +3,+5 & -5,+3 & -3,-5 \\ +1,+5 & -5,+1 & -1,-5 & +5,-1 \\ +5,+1 & -1,+5 & -5,-1 & +1,-5 \\ -3,+5 & -5,-3 & +3,-5 & +5,+3 \end{pmatrix} \end{array} \end{array},$$

for 64-QAM add:

$$\begin{array}{l} MOD(m_5 m_4 m_3 m_2 m_1 m_0) \qquad\qquad m_1 m_0 \\ \begin{array}{c} m_5 m_4 m_3 m_2 \\ 1\,1\,0\,0 \\ 1\,1\,0\,1 \\ 1\,1\,1\,1 \\ 1\,1\,1\,0 \\ 1\,0\,0\,0 \\ 1\,0\,0\,1 \\ 1\,0\,1\,1 \\ 1\,0\,1\,0 \end{array} \begin{array}{c} \begin{array}{cccc} 0\,0 & 0\,1 & 1\,1 & 1\,0 \end{array} \\ \begin{pmatrix} +5,+5 & -5,+5 & -5,-5 & +5,-5 \\ +5,-7 & +7,+5 & -5,+7 & -7,-5 \\ -7,-7 & +7,-7 & +7,+7 & -7,+7 \\ -7,+5 & -5,-7 & +7,-5 & +5,+7 \\ -3,-7 & +7,-3 & +3,+7 & -7,+3 \\ -7,+1 & -1,-7 & +7,-1 & +1,+7 \\ +1,-7 & +7,+1 & -1,+7 & -7,-1 \\ -7,-3 & +3,-7 & +7,+3 & -3,+7 \end{pmatrix} \end{array} \end{array}.$$

The outputs of the QPSK encoder form the least significant bits (LSBs), $m_1 m_0$, of the modulator input, and select the column of the matrix. The most significant bits (MSBs) determine the constellation size. With no uncoded bits ($m_5 = m_4 = m_3 = m_2 = 0$), QPSK is generated. With 2 uncoded bits, $m_3 m_2$, 16-QAM is generated. With 3 uncoded bits, $m_4 m_3 m_2$, 32-QAM is generated. With 4 uncoded bits, $m_5 m_4 m_3 m_2$, 64-QAM is generated. Furthermore, the effect of rotating the QAM constellation by 90° is to rotate the columns of the matrix,

$00 \rightarrow 01 \rightarrow 11 \rightarrow 10 \rightarrow 00;$

which leaves the rows invariant. This means the labeling of the uncoded bits is unaffected by 0°, 90°, 180° and 270° rotations. The handling of the 90° phase ambiguity at the receiver (decoder) is left solely to the QPSK encoder. Whatever method is used for resolving the ambiguity at the QPSK receiver can be directly incorporated into the QAM system using this labeling. For example, differential encoding of QPSK could be used if the QPSK code is itself rotationally invariant.

The labeling of a 16-QAM and 32-QAM constellation pattern in accordance with the present invention is illustrated in diagrammatic form in Figure 4. The constellation patterns, generally designated 80, correspond to the 16-QAM and 32-QAM matrices given above. In particular, for the 16-QAM example, the 16 constellation points are provided in a dashed box 90. The constellation points are divided into four subsets indicated by tokens 82, 84, 86, 88 as shown in Figure 5. Each subset contains four constellation points. Thus, for subset 82 designated by an open circle, four points 82a, 82b, 82c, and 82d are provided within box 90. The subset itself is defined by the two coded bits (QPSK bits) m0, m1 as indicated at 92 of Figure 6. For the 16-QAM implementation, the specific point within each subset is identified by the "uncoded" bits m2, m3 as indicated at 94 in Figure 6. Thus, 82c is defined as the 00 subset and the 011 point within that subset. Each remaining constellation point, such as points 84a, 86a, and 88a, are similarly identified.

For a 32-bit QAM implementation, the additional 16 points outside of dashed box 90 are also included. These points are labeled similarly, with all three bits m2, m3, m4 designated at 94 in Figure 6 being used. It will be appreciated that the labeling set forth can be expanded to higher levels of QAM.

A feature of the labeling scheme used in accordance with the present invention, as indicated in Figure 5, is that the Hamming weight of each QPSK symbol equals the Euclidian weight divided by a factor x, where x corresponds to the (minimum distance)^2 between constellation points. In the present example, the constellation points illustrated in Figure 4 are provided at QAM levels of 1, -1, 3, -3, 5, -5 in each of the quadrature channels, and therefore the minimum distance between constellation points is two, such that the Hamming weight is equal to the Euclidian weight divided by 4.

Figure 3 illustrates an implementation of a QAM trellis decoder in accordance with the present invention. The received symbol data is input to a pruner 62 via an input terminal 60. Pruner 62 processes the recovered modulation function to provide a set of metrics corresponding to the subsets defined by the QPSK codewords, and to provide a plurality of (N-2) bit subgroups representing a plurality of conditional determinations of the signal point identified by the transmitted uncoded bits. In particular, four metrics are output on line 66 to a rate 1/2 64-state Viterbi decoder 68. Four sets of (N-2) bit conditional determinations are output on line 64.

Pruner 62 can comprise a memory device, such as a programmable read only memory (PROM), that stores a look-up table containing precomputed sets of metrics and conditional determinations for different sets of input values $(\tilde{I}, \tilde{Q})$. The $(\tilde{I}, \tilde{Q})$ values are used to address the PROM to output the corresponding stored metrics and determinations. This allows a very high speed pruning operation. The Viterbi decoder uses an accumulated history of the metrics received from the pruner to decode the QPSK codewords.

The Viterbi decoder 68 illustrated in Figure 3 can be a conventional rate 1/2 decoder that is available for use with conventional QPSK coding schemes. Thus, in order to implement the decoder of the present invention, a custom Viterbi decoder is not required to decode the trellis codes.

Consider the process of signal detection when a soft-decision QPSK decoder is incorporated in a system employing the previously described QAM modulator. First, in hard-decision detection of QPSK or QAM signals, the received signal,

$$y_k = X_k + W_k,$$

is quantized, where the signal, x_k, belongs to the QPSK or QAM constellation (i.e., in the range of MOD(m)) and w_k is the noise. The quantization function produces an estimate of both the signal, $\hat{x}_k$, and the data $\hat{m}$, according to the relation, $\hat{x}_k = MOD(\hat{m})$. For maximum likelihood detection (ML), the log-likelihood function, $-\log(p(y_k \mid MOD(m)))$, is minimized over the possible messages, $m \in \{0, 1\}^N$, where $p(y_k \mid x_k)$ is the conditional probability of receiving y_k given that x_k is transmitted. For random messages, ML detection minimizes the probability of error. The most common method of quantization is nearest (Euclidean) neighbor detection, which satisfies

$$\|y_k - \hat{x}_k\|^2 = \min_{m \in \{0,1\}^n} \|y_k - \mathrm{MOD}(m)\|^2$$

where $\|.\|^2$ is the Euclidean distance squared (i.e., the sum of the squares). In the case of additive Gaussian noise, nearest neighbor detection is ML.

In coded QPSK and QAM systems, soft decision information should be provided to the decoder for effective decoding of the codeword. This soft-decision information is often described as a symbol metric; this metric indicates the quality of deciding a particular symbol, $\bar{x}_k = \mathrm{MOD}(\bar{m})$, was sent when y_k is received. For nearest neighbor decoding, the metric of choice is:

$$\mathrm{metric}(y_k; m) = \|y_k - \mathrm{MOD}(m)\|^2.$$

In practice, the metric itself is quantized for purposes of implementation. In QPSK, for example, for each possible message, $m_1, m_0 \in \{0, 1\}^2$, the nearest neighbor metric $\|y_k - \mathrm{MOD}(m_1, m_0)\|^2$ is the ML metric for additive Gaussian noise.

In trellis coded QAM modulation, based on a soft decision decodable QPSK code, four symbol metrics must be supplied to the decoder, as well as four conditional hard decisions. For nearest neighbor detection, for each choice of $m_1, m_0 \in \{0, 1\}^2$

$$\mathrm{metric}(y_k; m_1, m_0) = \min_{m_{N-1}, \cdots, m_2 \in \{0,1\}^{N-2}} \|y_k - \mathrm{MOD}(m_{N-1}, \cdots, m_2, m_1, m_0)\|^2;$$

the conditional hard decisions correspond to the choice of $m_{N-1}, \ldots, m_2$ that obtain the minimum. The process of determining the symbol metrics and conditional hard decisions is known as pruning. In trellis coded QAM, the uncoded bits appear as "parallel" branches of the trellis, and the computation of the symbol metrics and conditional hard decisions act to prune all but the single best branch from the set of parallel edges.

Note that pruning is easily described in terms of the QAM modulation matrix presented above. The pruning operation simply involves quantizing the received symbol, Y_k, for each column of the matrix. The conditional hard decision is then the best choice for each column and the metric corresponds to the quality of that decision.

Once the pruning operation has been completed, the soft decision information is presented to the decoder of the QPSK code. (During this time, the conditional hard decisions are stored waiting for the QPSK decisions.) The QPSK decoder, using the soft decision information, decodes the QPSK information (i.e., the m_1, m_0s). The remaining information (i.e., the $m_{N-1}, \ldots, m_2$s) is then decided in a well known manner using the decoded QPSK information and the previously stored conditional hard decisions.

Note that if the QPSK decoder is ML (for QPSK modulation) then the pruning/QPSK decoding method described is also ML. For example, if the QPSK code is a binary convolutional code with nearest neighbor (i.e., Viterbi) decoding, then the QAM trellis decoding algorithm is also nearest neighbor (i.e., finds the closest codeword to the received sequence).

In the embodiment illustrated in Figure 3, the metrics output from pruner 62 are decoded by decoder 68 to recover a single bit that corresponds to the single bit output on line 46 in the encoder of Figure 2. This bit is re-encoded with a rate 1/2 64-state convolutional encoder 70 (identical to encoder 48 in Figure 2) to recreate the two-bit QPSK codeword. The recreated codeword is used to select one of the four (N-2) bit subgroups output from the pruner, after the subgroups have been delayed by a delay buffer 72 for an amount of time equal to the delay introduced by decoder 68. The selected (N-2) bit subgroup is then combined with the recovered single bit from decoder 68 in a serializer 76, to provide a trellis decoded output.

As noted in connection with Figure 1, the decoded output may exhibit a modest symbol error rate that must be further improved by an outer decoder. Thus, further processing of the decoded output, by deinterleaver 34 and a Reed-Solomon outer decoder 36 (Figure 1) is used to recover the original information bits.

An estimate of the output bit-error rate, with a given input symbol error rate, for a t error-correcting, Reed-Solomon code can be easily computed. An (extended) Reed-Solomon code, over the finite field with $q = 2^l$, has parameters, (n_{RS} k, t), where the blocklength $n_{RS} \leq q + 1$, the dimension, $k = n_{RS} - 2t$, and the error-correction capability is t errors. For a memoryless, symbol error channel with input symbol error rate,

P_{in}, the output symbol error rate is bounded by:

$$P_{out} \leq (1/n_{RS}) \sum_{i=t+1}^{n_{RS}} \binom{n_{RS}}{i} (1-p_{in}^{\;n_{RS}-i}) \quad p_{in}^{\;i} \min(i + t,\ n_{RS}).$$

Then, the output bit error rate is approximated by the formula:

$P_b \approx P_{out} 2^{l-1}/(2^l-1)$

Also, if the l bit symbols of the Reed-Solomon code are composed of smaller, n bit symbols (e.g., the decoded symbols of a trellis coded QAM modulation) then the input error rate is:

$p_{in} \approx 1 - (1 - p_{mod})^{l/n}$

where p_{mod} is the n bit symbol error rate. To guarantee a "memoryless" channel when coded modulation is employed, the use of interleaving is required.

Figure 7 illustrates the performance of two concatenated systems, one employing conventional rate 2/3 trellis codes and decoding, and the other using the rate 1/2 QPSK implementation of trellis coded QAM in accordance with the present invention. The graph of Figure 7 plots Reed-Solomon block error rate against the carrier-to-noise ratio (CNR) in the received signal. A block error (or codeword error) occurs if one or more m-bit symbols are in error in the block. Curve 100 represents the performance of a concatenated Reed-Solomon trellis coded 16-QAM system in accordance with the present invention, using a rate 1/2, 64-state decoder. Curve 104 represents the performance of a similar system using trellis coded 32-QAM. Curve 102 represents the performance of a conventional trellis coded 16-QAM, rate 2/3, 16-state decoder. Curve 106 represents the performance of a conventional trellis coded 32-QAM rate 2/3, 16-state decoder.

The curves of Figure 7 were determined by using trellis coding simulation results to estimate the probability of an m-bit Reed-Solomon symbol being in error, P_{RSsym}, and then calculating the probability of a Reed-Solomon block error in accordance with the following formula:

$$P_{block} = \sum_{i=t+1}^{L} \binom{L}{i} P^{i}_{RSsym} (I-P_{RSsym})^{L-i}$$

where L is the Reed-Solomon block length (number of m-bit symbols per block) and t is the number of Reed-Solomon symbol errors that can be corrected per block. The 16-QAM system uses 116, 8-bit symbols per block, and the 32-QAM system uses 155, 8-bit symbols per block. Both Reed-Solomon codes can correct up to five, 8-bit Reed-Solomon symbols per block.

The curves in Figure 7 show that if it is desired or necessary to operate the system below a certain CNR, then the trellis coding approach of the present invention, represented by curves 100, 104, is clearly the correct choice. Even at higher CNRs, however, the trellis coding approach of the present invention may still be a better choice, because the trellis decoder apparatus can be produced in a more cost effective manner using a conventional QPSK Viterbi decoder chip.

It should now be appreciated that the present invention provides a practical system for digital transmission of power and band limited signals, such as compressed high definition television signals. A coded modulation scheme based on codes for QPSK modulation is directly incorporated into a QAM based modulation system, forming trellis coded QAM. This provides an easily implementable structure that is both efficient in bandwidth and data reliability.

Although the invention has been described in connection with specific embodiments thereof, those skilled in the art will appreciate that numerous adaptations and modifications may be made thereto without departing from the spirit and scope of the invention as set forth in the claims.

Claims

1. A method for communicating digital data using QAM transmission comprising the steps of:

dividing an N-bit QAM constellation pattern into four subsets, each subset including N/4 symbol points of said constellation pattern;

assigning a different two-bit codeword to each of said four subsets;

encoding a symbol to be transmitted by processing a first bit of said symbol with a rate 1/2 binary convolutional encoding algorithm to provide the two-bit codeword assigned to the subset in which said symbol resides in said constellation pattern;

mapping said two-bit codeword with the remaining bits of said symbol to provide a modulation function, wherein said remaining bits correlate said symbol with one of the N/4 symbol points included in the subset defined by said codeword; and

modulating a carrier with said modulation function for transmission on a communication channel.

2. A method in accordance with claim 1 wherein:

said two-bit codeword forms the least significant bits of said modulation function and defines the columns of a matrix of coordinates of said constellation pattern; and

said remaining bits form the most significant bits of said modulation function and determine the size of said constellation pattern.

3. A method in accordance with claim 1 or 2 comprising the further step of encoding information bits using an error correcting encoding algorithm to provide said symbol.

4. A method in accordance with claim 3 wherein said convolutional encoding step uses a trellis coding algorithm.

5. A method in accordance with any of the preceding claims comprising the further steps of:

receiving said carrier at a receiver;

demodulating the received carrier at said receiver to recover said modulation function;

pruning the recovered modulation function to provide a set of metrics corresponding to said subsets and to provide a plurality of bytes representing different conditional determinations of a signal point identified by the remaining bits;

using said metrics in an algorithm for decoding a rate 1/2 binary convolutional code to recover said first bit;

encoding the recovered first bit using a rate 1/2 binary convolutional encoding algorithm to recreate said codeword;

selecting one of said conditional determination bytes in response to said recreated codeword; and

combining said selected byte with the recovered first bit to provide a decoded output.

6. A method in accordance with claim 5 comprising the further steps of:

encoding information bits using an error correcting encoding algorithm to provide said symbol to be transmitted; and

further decoding said output using a symbol error correcting decoding algorithm.

7. A method in accordance with claim 5 or 6 wherein said algorithm for decoding is the Viterbi algorithm.

8. Apparatus for encoding digital data for QAM transmission comprising:

means for parsing a symbol to be transmitted into a first bit and at least one remaining bit;

means for encoding said first bit with a rate 1/2 binary convolutional encoding algorithm to provide a two-bit codeword that defines one of four subsets of an N-bit QAM constellation pattern, each subset including N/4 symbol points of said constellation pattern;

means for mapping said codeword with said remaining bits to provide a modulation function, wherein said remaining bits correlate said symbol with one of the N/4 symbol points included in the subset defined by said codeword; and

means for modulating a carrier with said modulation function for transmission on a communication channel.

9. Apparatus in accordance with claim 8 further comprising an outer encoder for encoding information bits using an error correcting encoding algorithm to provide said symbol.

10. Apparatus in accordance with claim 8 or 9 wherein:

said codeword forms the least significant bits of said modulation function and defines the columns of a matrix of coordinates of said constellation pattern; and

said remaining bits form the most significant bits of said modulation function and determine the size of said constellation pattern.

11. Apparatus in accordance with one of claims 8 to 10 wherein said encoding means use a trellis coding algorithm.

12. Apparatus for decoding QAM symbol data comprising:

means for demodulating a received carrier to recover an N-bit QAM modulation function in which a two-bit codeword identifies one of a plurality of QAM constellation subsets and the remaining (N-2) bit portion represents a signal point within said one subset;

means for pruning the recovered modulation function to provide a set of metrics corresponding to said subsets and to provide a plurality of (N-2) bit subgroups representing a plurality of conditional determinations of the signal point identified by the (N-2) bit portion;

decoder means for using said metrics in an algorithm for decoding a rate 1/2 binary convolutional code to recover a first bit;

means for encoding the recovered first bit using a rate 1/2 binary convolutional encoding algorithm to recreate said codeword;

means for selecting one of said plurality of (N-2) bit subgroups in response to said recreated codeword; and

means for combining the selected subgroup with the recovered first bit to provide a decoded output.

13. Apparatus in accordance with claim 12 wherein said codeword comprises the least significant bits in said modulation function and defines a column of a matrix of constellation coordinates, with the selected subgroup forming the most significant bits and defining a row of said matrix.

14. Apparatus in accordance with claim 12 or 13 wherein said pruning means quantize the recovered N-bit modulation function for each column of a matrix of constellation coordinates and said conditional determinations comprise a best choice for each of said columns with the set of metrics identifying the quality of each choice.

15. Apparatus in accordance with one of claims 12 to 14 wherein said decoding means comprise a decoder that uses a soft decision algorithm for decoding convolutional codes.

16. Apparatus in accordance with one of claims 12 to 15 further comprising:

an outer decoder for decoding said output using a symbol error correcting algorithm,

whereby the combination of said decoder means and said outer decoder form a concatenated decoder.

17. A concatenated decoder in accordance with one of claims 12 to 16 wherein said decoding algorithm comprises the Viterbi algorithm.

18. A concatenated decoder in accordance with claim 16 or 17 wherein said symbol error correcting algorithm comprises a Reed-Solomon code.

19. A concatenated decoder in accordance with one of claims 12 to 18 wherein said carrier is an HDTV carrier signal.

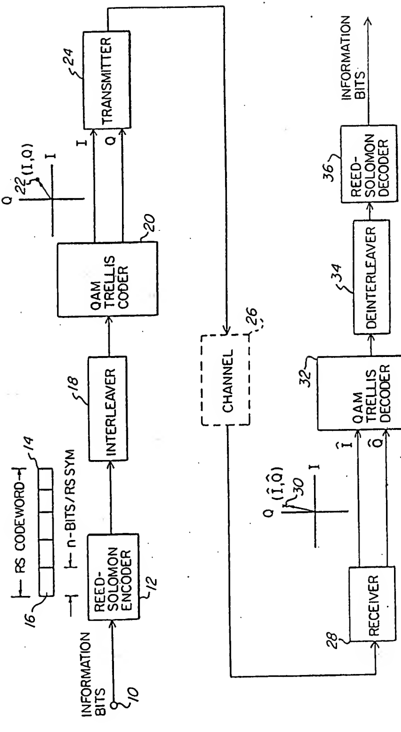

INFORMATION BITS
10
REED-SOLOMON ENCODER
12
n-BITS/RS SYM
RS CODEWORD
14
16
INTERLEAVER
18
QAM TRELLIS CODER
20
Q
I
(I,Q)
22
I
Q
TRANSMITTER
24
CHANNEL
26
RECEIVER
28
Q
I
(Î,Q̂)
30
Î
Q̂
QAM TRELLIS DECODER
32
DEINTERLEAVER
34
REED-SOLOMON DECODER
36
INFORMATION BITS

FIG. 1

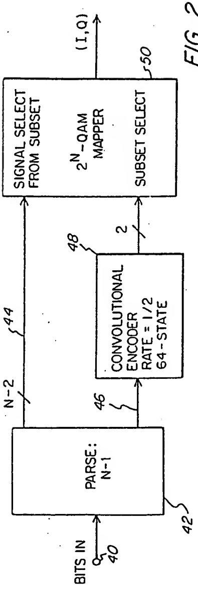

BITS IN
40
PARSE: N-1
42
N-2
44
46
CONVOLUTIONAL ENCODER RATE = 1/2 64-STATE
48
2
SIGNAL SELECT FROM SUBSET
2^N-QAM MAPPER
SUBSET SELECT
50
(I,Q)

FIG. 2

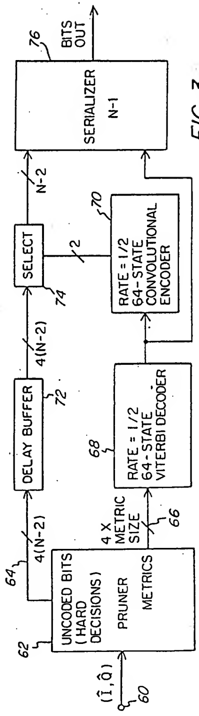

$(\hat{I},\hat{Q})$
60
62
UNCODED BITS (HARD DECISIONS)
PRUNER
METRICS
64
4(N-2)
4 X METRIC SIZE
66
DELAY BUFFER
72
RATE = 1/2 64-STATE VITERBI DECODER
68
4(N-2)
SELECT
74
2
RATE = 1/2 64-STATE CONVOLUTIONAL ENCODER
70
N-2
SERIALIZER N-1
76
BITS OUT

FIG. 3

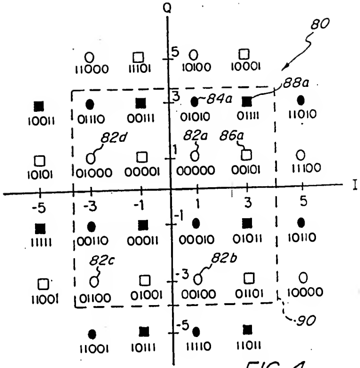
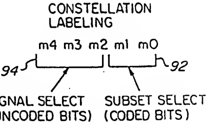


FIG. 4

	SUBSET LABELING			
	SUBSET	BINARY LABEL m1 m0	HAMMING WEIGHT	EUCLIDIAN WEIGHT/4
82	○	00	0	0
84	●	10	1	1
86	□	01	1	1
88	■	11	2	2

FIG. 5

FIG. 6



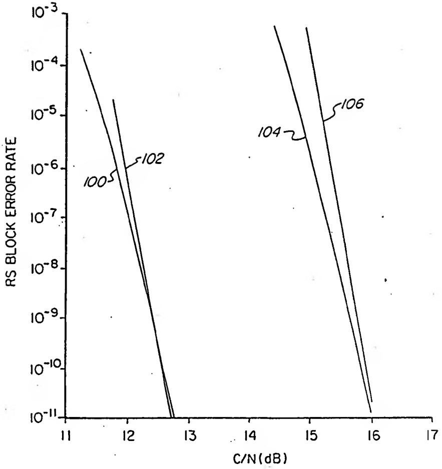

RS BLOCK ERROR RATE
10^-3
10^-4
10^-5
10^-6
10^-7
10^-8
10^-9
10^-10
10^-11
11
12
13
14
15
16
17
C/N(dB)
100
102
104
106

FIG. 7

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Cl

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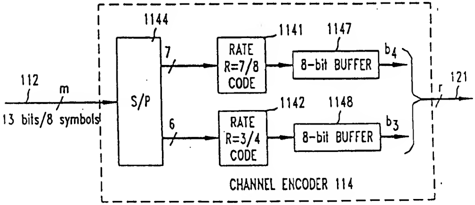
DOC

(54) Coded modulation with unequal error protection.

(57) Digital signals (from 101), such as digital television signals, are subjected to a source coding step (by 104) followed by a channel mapping step (by 114 and 115). The source coding step causes the signal to be represented by first and second data streams (on 105, 106). The first stream carries data regarded as more important and the second carries data regarded as less important. In the channel mapping step, the mapping is such that the data elements of the various data streams have differing probabilities of being erroneously detected at the receiver. The channel mapping step includes at least one multi-level coding step. The signal constellations used in the channel mapping step ere partitioned into supersymbols, in which the distance between the symbols comprising at least ones of the supersymbols is less than a parameter referred to as the maximum intra-subset distance (MID). Additionally, in some constellations, the minimum distance between at least ones of the symbols of at least one of the supersymbols is greater than the minimum distance between the symbols of the constellation as a whole while, again, still being less than the MID. The first data stream is used to identify a sequence of supersymbols, while the second data stream is used to select particular symbols from the identified supersymbols.

FIG. 5



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Background of the Invention

The present invention relates to the transmission of digital data including, particularly, the transmission of digital data which represents television (TV) signals.

It is generally acknowledged that some form of digital transmission will be required for the next generation of television technology, conventionally referred to as high definition television, or HDTV. This requirement is due mostly to the fact that much more powerful video compression schemes can be implemented with digital signal processing than with analog signal processing. However, there has been some concern about getting committed to an all-digital transmission system because of the potential sensitivity of digital transmission to small variations in signal-to-noise ratio, or SNR, at the various receiving locations.

This phenomenon--sometimes referred to as the "threshold effect"--can be illustrated by considering the case of two television receivers that are respectively located at 50 and 63 miles from a television broadcast station. Since the power of the broadcast signal varies roughly as the inverse square of the distance, it is easily verified that the difference in the amount of signal power received by the television receivers is about 2 dB. Assume, now, that a digital transmission scheme is used and that transmission to the receiver that is 50 miles distant exhibits a bit-error rate of 10^{-6}. If the 2 dB of additional signal loss for the other TV set translates into a 2 dB decrease of the SNR at the input of the receiver, then this receiver will operate with a bit-error rate of about 10^{-4}. With these kinds of bit-error rates, the TV set that is 50 miles away would have a very good reception, whereas reception for the other TV set would probably be very poor. This kind of quick degradation in performance over short distances is generally not considered acceptable by the broadcasting industry. (By comparison, the degradation in performance for presently used analog TV transmission schemes is much more graceful.)

There is thus required a digital transmission scheme adaptable for use in television applications which overcomes this problem. Solutions used in other digital transmission environments--such as the use of a) regenerative repeaters in cable-based transmission systems or b) fall-back data rates or conditioned telephone lines in voiceband data applications--are clearly inapplicable to the free-space broadcast environment of television.

An advantageous technique for overcoming the shortcomings of standard digital transmission for over-the-air broadcasting of digital TV signals-- referred to herein generically as "unequal-error-protection signaling"--comprises a particular type of source coding step followed by a particular type of channel mapping step. More specifically, the source coding step causes the television signal to be represented by two or more data streams while, in the channel mapping step, the mapping is such that the data elements of the various data streams have differing probabilities of channel-induced error, i.e., different probabilities of being erroneously detected at the receiver. Illustratively, a first one of the aforementioned data streams carries components of the overall television signal which are regarded as the most important--for example, the audio, the framing information, and the vital potions of the video information--and that data stream is mapped such that its data elements have the lowest probability of channel-induced error. A second one of the data streams carries components of the overall television signal which are regarded as less important than those of the first data stream and that data stream is mapped such that its data elements have a probability of channel-induced error that is not as low as those of the first data stream. In general, it is possible to represent the overall television signal with any number of data streams, each carrying components of varying importance and each having a respective probability of error. This approach allows a graceful degradation in reception quality at the TV set location because, as the bit error rate at the receiver begins to increase with increasing distance from the broadcast transmitter, it will be the bits that represent relatively less important portions of the TV signal information that will be the first to be affected.

In a scheme which implements the above-described overall concept in which different levels of error protection are provided for different classes of data elements generated by the source enclosing step, but which provides enhanced noise immunity via the use of coded modulation, such as trellis-coded modulation, the symbols in a predetermined 2N-dimensional channel symbol constellation, $N \geq 1$, are divided into groups, each of which is referred to as a "supersymbol." During each of a succession of symbol intervals, a predetermined number of the most important data elements are channel encoded, and the resulting channel coded data elements identify a particular one of the supersymbols. The remaining data elements, which may also be channel encoded, are used to select for transmission a particular symbol from the identified supersymbol.

The approaches thus far described is similar in a general way to conventional coded modulation schemes in that the latter also divide the channel symbols into groups, typically referred to as "subsets." However, in conventional coded modulation schemes, the subsets are formed under the constraint that the minimum Euclidean distance (hereinafter referred to as the "minimum distance") between the symbols in a subset is greater than the minimum distance between the symbols in the constellation as a whole. In the described approach, however, the minimum distance between the symbols of a supersymbol is the same as the minimum distance

between the symbols in the constellation as a whole. This distance property allows for greater amount of noise immunity for the most important data elements than for the other data elements, that immunity being optimized by keeping the minimum distance between supersymbols as large as possible--usually greater than the minimum distance between the symbols of the constellation. Specifically, once the supersymbols are defined, it is possible to design codes for the most important data elements based on the distances between the supersymbols, i.e., as though each supersymbol were a conventional symbol in a conventional constellation. This being so, a particular degree of noise immunity can be achieved for the most important data elements that is greater than what can be achieved for the other data elements.

Summary of the Invention

The present invention provides the designer of unequal-error-protection signaling schemes of the above-described type with additional flexibility. In accordance with the invention, so-called multi-level coding is used to code the data elements of at least one of the data streams that are input to the channel mapping step. For example, multi-level coding can be used to code the data elements which ultimately determine the supersymbol selection. Alternatively, it may be used to code the data elements which determine the selection of a particular symbol from within a selected supersymbol. Or multi-level coding can be used for both streams. The particular way in which the multi-level coding is used will depend on the requirements of any particular application in terms of the degree of error protection desired to be afforded to any particular class of the data being coded.

There are at least two important advantages to this approach. One is that it provides an enhanced flexibility in designing a channel coding to realize a desired percentage of the overall data stream being coded that is to be regarded, and treated, as most important. Another advantage is that it provides an enchanced flexibility in apportioning the available redundancy between the most important data and the less important data, thereby providing a mechanism for achieving particular desired different levels of error protection for those two classes of data. A yet further advantage is that this approach allows differential levels of protection to be afforded to substreams of data elements within any of the streams that are multi-level coded in accordance with the invention.

Multi-level coding per se is a technique already known in the prior art. In accordance with that technique, data elements to be coded are divided into two or more substreams. Each of one or more of the substreams is then individually redundancy coded using a code of any desired type. The individual encoded substreams--along with any substreams that were left uncoded--form the output of the multi-level code. That output is then used in the prior art to identify channel symbols of a predetermined constellation for transmission over a channel. However, the prior art does not encompass the teaching, lying at the heart of the present invention, that multi-level coding can be advantageously used to code one or more of the data streams of an overall unequal-error-protection signaling scheme.

Advantageously, particular desired combinations of a) coding gain for the most important data elements, b) coding gain for the less important data elements, and c) percentage of most important data elements are more readily achievable by incorporating one or more multi-level codes in an unequal-error-protection signaling scheme in accordance with the invention, than when single-level codes are used. From the coding theory standpoint, this result can be understood as arising out of the fact that the invention allows the redundancy introduced into the overall coding scheme to be allocated in virtually any desired proportion between the coding of the most important data elements and coding of the less important data elements.

Brief Description of the Drawing

FIG. 1 is a block diagram of a transmitter embodying the principles of the invention;
FIG. 2 is a block diagram of a receiver for signals transmitted by the transmitter of FIG. 1;
FIG. 3 depicts a prior art signal constellation;
FIG. 4 depicts a signal constellation illustratively used by the transmitter of FIG. 1;
FIGS. 5 and 6 show illustrative multi-level coders used in the transmitter of FIG. 1 in accordance with the invention;
FIG. 7 depicts a signal constellation that can alternatively be used by the transmitter of FIG. 1;
FIG. 8 depicts a signal constellation of the type typically used in equal error protection schemes;
FIGS. 9-14 depict further signal constellations that can alternatively be used by the transmitter of FIG. 1; and
FIGS. 15 and 16 show illustrative multi-stage decodes used in the receiver of FIG. 2 in accordance with the invention.

Detailed Description

Before proceeding with a description of the illustrative embodiments, it should be noted that various ones of the digital signaling concepts described herein are all well known in, for example, the digital radio and voiceband data transmission (modem) arts and thus need not be described in detail herein. These include such concepts as multidimensional signaling using 2N-dimensional channel symbol constellations, where N is some integer, trellis coding; scrambling; passband shaping; equalization; Viterbi, or maximum-likelihood, decoding; etc. These concepts are described in such U.S. patents as U.S. 3,810,021, issued May 7, 1974 to 1. Kalet et al.; U.S. 4,015,222, issued March 29, 1977 to J. Werner; U.S. 4,170,764, issued October 9, 1979 to J. Salz et al.; U.S. 4,247,940, issued January 27, 1981 to K. H. Mueller et al.; U.S. 4,304,962, issued December 8, 1981 to R. D. Fracassi et al.; U.S. 4,457,004, issued June 26, 1984 to A. Gersho et el.; U.S. 4,489,418, issued December 18, 1984 to J. E. Mazo; U.S. 4,520,490, issued May 28, 1985 to L. Wei; and U.S. 4,597,090, issued June 24, 1986 to G. D. Forney, Jr.

It may also be noted before proceeding that various signal leads shown in the FIGS. may carry analog signals, serial bits or parallel bits, as will be clear from the context.

Turning now to FIG. 1, television (TV) signal source 101 generetes an analog video signal representing picture information or "intelligence," which signal is passed on to source encoder 104. The latter generates a digital signal comprised of data elements in which at least one subset of the data elements represents a portion of the information, or intelligence, that is more important than the portion of the informetion, or intelligence, represented by the rest of the data elements. Illustratively, each date element is e data bit, with an average m+k information bits being generated for each of a succession of symbol intervals. The symbol intervals are comprised of N signaling intervals, where 2N is the number of dimensione of the constelletion (as described below). The signaling intervals have duration of T seconds and, accordingly, the symbol intervals each have a duration of NT seconds. The embodiments explicitly disclosed herein heppen to use two-dimensional constellations, i.e., N = 1. For those embodiments, then, the signaling intervals and the symbol intervals are the same.

Of the aforementioned m+k information bits, the bits within the stream of m bits per symbol interval, appearing on lead 105, are more important then the bits within the stream of k bits per symbol interval, appearing on lead 106.

The bits on leads 105 and 106 are independently scrambled in scramblers 110 and 111, which respectively output m end k parallel bits on leads 112 and 113. (Scrambling is customarily carried out on a serial bit stream. Thus although not explicitly shown in FIG. 1, scramblers 110 and 111 may be assumed to perform a parallel-to-serial conversion on their respective input bits prior to scrambling and a serial-to-parallel conversion et their outputs.) The signal is then channel mapped. In particular, the respective groups of bits on leads 112 and 113 are extended to channel encoders 114 and 115 which generate, for each symbol interval, respective expanded groups of r and p bits on leads 121 and 122, where r > m and p > k. The values of those bits jointly identify a particular chennel symbol of a predetermined constellation of channel symbols (such as the constellation of FIG. 4 as described in detail hereinbelow). Complex plane coordinates of the identified channel symbol are output by constellation mapper 131, illustratively realized as a lookup table or as a straightforward combination of logic elements. Conventional pass-band shaping and television modulation are then performed by passband shaper 141 and television modulator 151, respectively. The resultant analog signal is then broadcast via antenna 152 over a communication channel, in this case a free-space channel.

In order to understand the theoretical underpinnings of the invention, it will be useful at this point to digress to a consideration of FIG. 3. The latter depicts a standard two-dimensional data transmission constellation of the type conventionally used in digital radio and voiceband data transmission systems. In this standard scheme--conventionally referred to as quadrature-amplitude modulation (QAM)--data words each comprised of four information bits are each mapped into one of 16 possible two-dimensional channel symbols. Each channel symbol has an in-phase, or I, coordinate on the horizontal axis and hes e quadrature-phase, or Q, coordinate on the vertical axis. On each axis, the channel symbol coordinates are ± 1 or ± 3 so that the distance between each symbol and each of the symbols that are horizontally or vertically adjacent to it is the same for all symbols --that distance being "2". As a result of this uniform spacing, essentially the same amount of noise immunity is provided for all four information bits.

As is well known, it is possible to provide improved noise immunity without sacrificing bandwidth efficiency (information bits per signaling interval) using a coded modulation approach in which an "expanded" two-dimensional constellation having more than (in this example) 16 symbols is used in conjunction with a trellis or other channel code. For example, one can use a 32-symbol, two-dimensional constellation together with an 8-state trellis code to achieve approximately 4 dB of enhanced noise immunity over the uncoded case of FIG. 3, while still providing for the transmission of four information bits per signaling interval. Here, too, however,

essentially the same amount of noise immunity is provided for all four information bits.

Moreover, it is known that the known noise immunity and bandwidth efficiency advantages of coded modulation are achieved while providing different levels of protection against channel-induced error for different classes of bits. Specifically, it is possible to achieve a level of error protection for a class of most important bits which is substantially greater than what can be achieved with the aforementioned conventional coded modulation approach. Indeed, the transmitter of FIG. 1 embodies that concept, as will now be described in further detail.

The constellation used in the transmitter of FIG. 1 is illustratively the two-dimensional 32-symbol constellation shown in FIG. 4. The symbols in the signal constellation are divided into groups referred to as "supersymbols." Specifically, the constellation of FIG. 4 is divided into $2^r = 2^2 = 4$ supersymbols. In this example, the points in the four quadrants constitute respective supersymbols, as denoted by a box enclosing each group. The supersymbols are denoted generically as $\Omega_{b_4b_3}$ where b_4 = 0,1 and b_3 = 0,1. The four supersymbols are thus Ω_{00}, Ω_{01}, Ω_{10}, and Ω_{11}.

In this example, m = 1.625 and k = 2.125 so that the overall bit rate is 3.75 bits per symbol with 43.33% of the bits being in the class of most important bits. (The manner in which such fractional average bit rates can be achieved in practice will become clear as this description continues.) Encoder 114 adds an average 0.375 redundant bits for every 1.625 bits that are input to encoder 114 so that r = 2. Encoder 115 adds an average 0.875 redundant bits for every 2.125 bits that are input to encoder 115 so that p = 3. The r = 2 bits on lead 121 identify one of the four supersymbols and the p = 3 symbols on lead 122 select a particular one of the eight channel symbols within the identified supersymbol. The partitioning of the constellation is such that the minimum distance between the symbols of a supersymbol--that distance being denoted as d_2 in FIG. 4--is the same as the minimum distance between the symbols in the constellation as a whole. Given this characteristic, increased noise immunity for the most important bits can be provided via appropriate selection of a) the codes implemented by encoders 114 and 115 and b) the ratio d_1/d_2--where d_1 is the minimum distance between the supersymbols. (The parameter d_1 is given by the minimum of the distances between all the pairs of supersymbols, with the distance between any pair of supersymbols being the minimum distance between any symbol of one of the pair of supersymbols and any symbol of the other.)

Specifically, a coded modulation scheme can now be constructed for the most important bits as though the four supersymbols were four conventional symbols in a conventional constellation. To design such a coded modulation scheme, the four supersymbols are partitioned, as is conventional, into a predetermined number of subsets. In this case, there are two subsets, where the value of b_3 denotes which subset each of the supersymbols belongs to. Thus, one subset, referred to as subset "0", comprises supersymbols Ω_{00} and Ω_{10} and the other, referred to as subset "1", comprises supersymbols Ω_{01} and Ω_{11}. An appropriate code is used to encode the most important input bits to generate a stream of coded output bits which a) define a sequence of these subsets and b) for each subset of the sequence, select a particular supersymbol within that subset. The less important bits are then used to select for transmission a particular symbol from each of the thus-selected supersymbols. In this example, as already seen, this latter selection also involves the use of coded modulation.

<u>In accordance with the present invention</u>, at least one of the channel encoders implements a multi-level code. <u>In this example, in particular, both of</u> them do. As noted earlier, a multi-level code is one in which the data elements--in this embodiment, bits--to be coded are divided into two or more substreams. Each of one or more of the substreams is then individually redundancy coded using a code of any desired type. The individual encoded substreams--along with any substreams that were left uncoded--form the output of the multi-level code.

Particular illustrative embodiments for encoders 114 and 115 are shown in FIGS. 5 and 6, respectively. Encoder 114 implements a two-level code and, as such, includes two coders--coders 1141 and 1142. The redundancy code implemented by coder 1141 is a conventional rate R = 7/8 zero-sum parity check code such as shown in G. C. Clark, Jr. and J. B. Cain, <u>Error-Correction Coding for</u> Digital <u>Communications</u> New York: Plenum, 1981). The redundancy code implemented by coder 1142 is a conventional rate R = 3/4 punctured convolutional code such as shown in Y. Yasuda et al, "High-rate punctured convolutional codes for soft decision Viterbi decoding," <u>IEEE Trans. Comm.</u>, Vol. COM-32, 1984, pp. 315-318. In operation, a serial-to-parallel (S/P) converter 1144 within encoder 114 takes in 13 bits over 8 symbol intervals, yielding an average input bit rate of m =1.625 bits per symbol interval as noted earlier. The output of converter 1144 comprises two substreams of bits. In one substream, the bits are provided in parallel groups of seven to coder 1141. In the other substream, the bits are provided in parallel groups of six to coder 1142. For every group of seven input bits, coder 1141 generates eight output bits which are applied to 8-bit buffer 1147. At the same time, for every group of six input bits, coder 1142 generates eight output bits which are applied to 8-bit buffer 1148. The contents of buffers 1147 and 1148 are read out synchronously in such a way that a pair of bits--one from each of the two buffers--is

provided on lead 121 for each symbol interval. These bits are the aforementioned bits b_3 and b_4. As such, the former identifies one of the two subsets "0" and "1" and the other identifies one of the two supersymbols from the identified subset, the two bits b_3 and b_4 thus identifying one of the four supersymbols Ω_{00}, Ω_{01}, Ω_{10} and Ω_{11}.

Encoder 115 implements a three-level code and, as such, includes coders 1151, 1152 and 1153. Coders 1151 and 1152 implement the same codes that are implemented by coders 1141 and 1142, respectively. Coder 1153 implements a rate R = 1/2 convolutional code such as shown in the aforementioned Clark and Cain text. In operation, a serial-to-parallel (S/P) converter 1154 takes in 17 bits over 8 symbol intervals, yielding an average input bit rate of k = 2.125 bits per symbol interval, as noted earlier. The output of converter 1154 comprises three substreams of bits. In one substream, the bits are provided in parallel groups of seven to coder 1151. In the second stream, the bits are provided in parallel groups of six to coder 1152. In the third group, the bits are provided in parallel groups of four to coder 1153. For every group of seven input bits, coder 1151 generates eight output bits which are applied to 8-bit buffer 1157. For every group of six input bits, coder 1152 generetes eight output bits which are applied to 8-bit buffer 1158. For every group of four input bits, coder 1153 generates eight output bits which are applied to 8-bit buffer 1159.

The contents of buffers 1157, 1158 and 1159 are read out synchronously in such a way that three bits--one from eech of the three buffers--are provided on lead 122 for eech symbol interval. These three bits--denoted b_2, b_1 and b_0-- select a particular symbol from the supersymbol identified at the output of encoder 114. To this end, each of the symbols in the constellation is labelled with a three-bit label as shown in FIG. 4. These three bits are, in fact, the eforementioned bits b_2, b_1 and b_0.

The symbols of each supersymbol are partitioned at a first level of partitioning into two subsets. Each subset is comprised of four symbols as denoted by the labelled b_0 value. Thus one subset is comprised of the four symbols denoted 000, 010, 100 and 110 and the other subset is comprised of the four symbols denoted 001, 011, 101 and 111. The symbols of each of these subsets are partitioned at a second level of pertitioning into two second-level subsets, which ere identified by their labelled b_0 and b_1 values. Each second-level subset is comprised of two symbols identified by b_2.

The assignment of three-bit labels to the symbols of each supersymbol is not erbitrary. Rather, the codes implemented by coders 1151, 1152 and 1153 are selected taking into eccount the minimum distance between the subsets at each level of pertitioning. In particuler, the most powerful, i.e., lowest-rate, code--in this case the rate R = 1/2 code implemented by coder 1153--is used et the first level of partitioning to generate b_0 because the minimum distance between the subsets at the first level is the smallest, that distence being d_1. The second- and least-powerful codes, the rate R = 3/4 and rate R = 7/8 codes implemented by coders 1152 and 1151, respectively, generate b_1 and b_2, respectively, because the minimum distance between the second-level subsets is greater by a factor of $\sqrt{2}$ than that of the first level minimum distance whereas the minimum distance between the symbols in each second-level subset is greater by a factor of 2 than the first-level minimum distance.

The advantages provided by such use of multi-level codes in an unequal error protection signaling scheme are discussed hereinbelow. First, however, reference is made to the receiver of FIG. 2.

In particular, the analog broedcast signal is received by antenna 201; subjected to conventional television front-end processing in processing unit 211 including, for example, demodulation; and converted to digital form by A/D converter 212. The signal is then equalized by passband channel equalizer 221, which generates a signal representing the equalizer's best estimate es to the I and Q component values of the transmitted symbol. This estimate, which is referred to hereinbelow as the "received symbol signal," is passed on parallel rails 222 and 223 to be channel decoded by channel decoders 231 and 232. The function of channel decoder 231 is to identify the most likely sequence of supersymbols, while the function of channel decoder 232 is to identify the most likely sequence of symbols, given that sequence of supersymbols. The output of decoder 232 on lead 234 thus comprises the bits b_3 and b_4, while the output of decoder 231 on lead 233 thus comprises the bits b_0, b_1 and b_2.

Since the streams of most-important and less-important bits are, in this embodiment, both multi-level coded, the channel decoders 231 and 232 must each be multi-level decoders. Streightforward maximum likelihood decoding, such as a Viterbi decoding, could be employed for this purpose. However, the present illustrative embodiment uses a more refined approach to multi-level decoding--an approach referred to es multi-stage decoding. This is a well-known technique, the details of which can be found in A. R. Calderbank, "Multilevel codes and multistage decoding," IEEE Trans. Comm., Vol. COM-37, pp. 222-29, 1989, hereby incorporated by reference. For present purposes, then, it suffices to summarize, in overview, how the multi-stage decoding is carried out.

In particular, channel decoder 232 first recovers the bits encoded by coder 1142 within encoder 114 (FIG. 5), independent of, and without reference to, any decoding performed on the bits encoded by coder 1141. To

this end, as shown in FIG. 15, decoder 232 includes decode b_3 circuitry 2321, decode b_4 circuitry 2323 and delay element 2322. In operation, the received symbol signal is processed by circuitry 2321 to decode bit b_3 by first finding the symbol closest to the received signal symbol--and the associated metrics--in (Ω_{01} v Ω_{11}) and in (Ω_{00} v Ω_{10}). The trellis path is then extended as per conventional Viterbi decoding of the code implemented by coder 1142 (FIG. 5) to generate a final decision about a previously encoded information bit that was encoded by that coder. The decoded bit--one of the original most important bits--is provided as an output on lead 234. It is also re-encoded within circuitry 2321 using the code of coder 1142 to provide bit b_3 to circuitry 2323 as well as to channel decoder 231 (FIG. 16), as described below, via lead 236. A version of the received symbol signal is delayed by delay element 2322 by a time sufficient to provide the value of bit b_3 as just described. That signal is then provided to circuitry 2323 along with bit b_3 to circuitry 2323 which then proceeds to decode bit b_4. This is carried out, in particular, by first finding the symbol closest to the received signal symbol--and the associated metrics--in supersymbol Ω_{0b_3} and in supersymbol Ω_{1b_3}. The metrics are stored in a buffer and are used to perform maximum likelihood decoding of the code implemented by clear 1141 (FIG. 5), thereby to provide another of the most important bits on lead 234. At the same time, that bit is re-encoded within circuitry 2323 to provide bit b_4 to channel decoder 231 on lead 236.

It may be noted at this point that, obviously, circuitries 2321 and 2323 must be provided within information about the constellation that is being used in order to carry out their respective functions. That information is illustratively stored in a constellation store 2325 whose output--denoted as "A"--is provided to both of those circuitries, as well as to decoder 231 of FIG. 16.

With the most important bits now provided on leads 234 and the values of bits b_3 and b_4 now being provided to decoder 231 on lead 236, the multi-stage decoding can now proceed--within the latter channel decoder--to recover the less important bits. To this end, as shown in FIG. 16, decoder 231 includes decode b_0 circuitry 2315, decode b_1 circuitry 2316, decode b_2 circuitry 2317, and delay elements 2311, 2312 and 2313. In operation, the received symbol signal is delayed by an amount equal to the processing delay within decoder 232 so that circuitry 2315 can be provided with bits b_3 and b_4 at the same time that it receives the received symbol signal. Circuitry 2315 then proceeds by first finding, within the supersymbol $\Omega_{b_4b_3}$, those symbols having b_0 = 0 and b_1 = 1 which are respectively closest to the received signal symbol. The trellis path is then extended, as before, using the metrics associated with those two closest symbols, leading ultimately to a decoding of bit b_0 and hence a recovery of one of the less important bits. That bit is then re-encoded within circuitry 2315 so as to be able to provide the value of bit b_0 to circuitries 2316 and 2317. The latter operate in a manner similar to that described above with respect to the other decode circuitries to recover the other less important bits, with the delays of delay elements 2312 and 2313 being sufficient to enable each of the circuitries 2316 and 2317 to receive the re-encoded bits, as needed.

Decoding in the case where multi-dimensional symbols are used is carried out in a similar manner, as will be appreciated by those skilled in the art.

The bits output by decoder 231 and 232 on leads 233 and 234, respectively, are descrambled by descramblers 241 and 242, which respectively perform the inverse function of scramblers 110 and 111 in the transmitter. A television signal formatted so as to be displayable by an appropriate television set is then generated from the descrambler outputs by source decoder 253, thereby recovering the original television information, or intelligence. That signal is then presented to the television viewer on television set 260.

The performance of the overall unequal error protection signaling scheme implemented by the system of FIGS. 1 and 2, as just described, can now be characterized in terms of the nominal coding gain (i.e., the coding gain at very low error rates), this being the gain in signal-to-noise ratio over that of an uncoded 16-QAM system. The gain for the most important bits is 6.24 dB and for the less important bits is 2.70 dB when the number of states of each of the convolutional codes is chosen to be 16. The unique advantage of the present invention, however, does not wholly lie with the particular levels of coding gain achieved. Coding arrangements known in the prior art may achieve comparable, or even better, coding gain results in particular applications. However, where the invention is particularly advantageous is in its ability to provide the system designer with a greatly enhanced capability to pick the desired performance criteria, given the application, and to then readily arrive at a coding scheme which meets those criteria.

For example, the overall data rate of the above-described system can be increased from 3.75 to 3.875 bits per symbol without affecting the coding gain by changing the code implemented by coders 1141 and 1151 to a rate R = 15/16 zero-sum parity check code (and using 16-bit, rather than 8-bit, buffers in the encoders). (At the same time, the percentage of most important bits increases ever so slightly-- from 43.33% to 43.55%.) As another example, the percentage of most important bits can be reduced to 34.375%, while at the same time increasing the level of error protection afforded to those bits by a) changing the codes implemented by coders 1141, 1142, 1152 and 1153, to respectively, a rate R = 7/8 punctured convolutional code; a rate R = 3/4

convolutional code; a rate R = 7/8 zero-sum parity check code; end a rete R = 3/4 convolutional code; and b) eliminating coder 1151 so that the bits applied to buffer 1157 are uncoded bits. This arrangement achieves coding gains of 8 dB for the most important bits and 0.22 dB for the less important bits. Additionally, by changing the ratio of d_2/d_1 in FIG. 4, one can trade off the gains of the most important and less important bits. It will thus be seen that the invention allows for the use of a virtually unlimited range of design parameters-- including the code rates, code complexity, overall coding redundancy, fraction of that redundancy used for error protection of the most important bits (as opposed to that used for the less important bits)--in order to meet desired system design criteria. This flexibility can be further enhanced by using various different signal constelletions, including constelletions having various different numbers of symbols, symbol spacings, supersymbol groupings and subset partitionings. Indeed a new class of constellations may be advantegeously used to provide the system designer with even further flexibility. These constellations are characterized by particular distance relationships within the constellation. Indeed, an important parameter in the design of coded modulation schemes is the so-called intra-subset distance. This perameter is the minimum distance between any two symbols in the subset. In coded modulation schemes which, unlike in the present invention, seek to provide equal error protection, the design constraint is to pertition the constellation into subsets in such e way as to meximize the minimum of the intra-subset distances taken ecross all the subsets. This value, which we define as the "maximum intra-subset distance," or MID, has been achieved when, given a particular partitioning, any further attempt to increase the intra-subset distance of any particular subset (by making eny other symbol-to-subset assignment changes) would not result in a further increase in the aforesaid minimum value.

Further, an important distinction between equal error protection and unequal error protection schemes needs to be kept in mind In the former, the error protection for the so-called coded bits is determined by the minimum distance between subset sequences while the error protection for the uncoded bits is determined by the minimum distance between the symbols within a subset. Designers of equal-error-protection schemes want these two minimums to be as close to each other as possible because they went equal error protection for all the data. The performance of these schemes is dominated by the distance between the symbols within a subset. This results from the fact that one can always increase the complexity of the code in order to achieve whatever distance is desired between subset sequences.

In the case of unequal error protection, by contrast, the symbols in a supersymbol are selected by the less important bits. Thus the distance between these symbols can be significently less then the MID. Indeed it is limited only by the distance between the symbols of the constellation es a whole. This distence is chosen to provide the necessary level of error protection for the less important bits. Once we fix that distance, then the subset partitioning cen be done to take advantage of this fact, thus making possible the realization of greeter distances between the supersymbol sequences than would be possible with conventional coded moduletion for the same complexity. We are, thus, no longer constrained to keep the symbols within a supersymbol far-eway from each other.

Taking the foregoing into account, constellations useful in unequal-error-protection signaling schemes are characterized by the fact that the minimum distance between at least some of the symbols of at least one of the supersymbols is less than the MID. Indeed, constellations of this generel type are known. However, in the prior ert, the minimum distence between the symbols of the supersymbols is the same es the minimum distance between the symbols of the constellation as a whole. By contrast, the constellations in question are not so constrained. That is, the minimum distance between et leest ones of the symbols of at least one of the supersymbols is greater than the minimum distance between the symbols of the constellation as a whole while, egain, still being less than the MID. Graphically speeking, constellations meeting this criterion will generally appear to have supersymbols which, at least to some degree, overlap (as do the subsets of the constellations used for equal error protection schemes). That is, at leest one symbol of a leest one supersymbol will be closer to each of e pair of symbols of a different supersymbol than that pair of symbols are to each other.

One illustrative, 32-symbol, constellation embodying these principles is shown in FIG. 7. This constellation is partitioned into four supersymbols, the constituent symbols thereof being labelled A, B, C and D, respectively. The distance labelled as "X" can be, for exemple, "1", thereby providing e constelletion whose individual symbols are uniformly spaced. Or "X" can be greater than one--such as $\sqrt{3}$--thereby increasing some of the inter-supersymbol distances. This provides an additional degree of design freedom in terms of achieving particular desired levels of error protection.

The MID of the constellation of FIG. 7 is "2", which can be verified from a consideration of FIG. 8, in which the same constelletion has been partitioned in order to provide equal error protection. That is, the minimum distance between any two symbols labelled "A", for example, is, in fact, "2". In FIG. 7, by contrast, the corresponding minimum distance, in accordence with the invention, is less than "2". Specifically, it is $\sqrt{2}$. And note that, from a graphical perspective, subsets A and B overlap one another, as do subsets C and D. (It is thus not possible to draw a box around the supersymbols, es was done in FIG. 4.)

Another illustrative constellation--this one having 64 symbols--is shown in FIG. 9. Again, the constellation is partitioned into four supersymbols labelled using the same labelling convention as that used for FIG. 7.

FIGS. 10-14 show yet further constellations that can be used to advantage in conjunction with the present invention. The same labelling conventions are used and thus nothing further need be said about the configuration of these constellations.

The following two tables illustrate the tremendous flexibility provided by the present invention in terms of providing unequal-error-protection signaling schemes which provide various different combinations of a) percentage of most important bits, b) degree of overall code redundancy, c) coding gains, and d) ratio of peak-to-average power (which is an important consideration in power-limited channels such as terrestrial and satellite telephone channels). TABLE 1 is a table of codes in which 25% of the bits constitute the most important bits. TABLE 2 is a table of codes in which various other percentages of the bits constitute the most important bits.

TABLE 1

Example	Signal Constellation	Code C_0 for important bits	Code C_1 for less important bits
4A 25% most important data rate $4-1/L$	Fig. 10 $P = 5/2,\ PAR = 1{\cdot}7$ $d^2(\Omega_{ab},\Omega_{cd}) \geq \begin{cases} 4, & \text{if } ab = \bar{c}\bar{d} \\ 1, & \text{if } ab \neq cd \end{cases}$	[23,35]: 16-state, rate 1/2 convolutional code $d^2(C_0) = 11$ $\Gamma = 7.4$ db	single parity check code with redundancy $1/L$ $d^2(C_1) = 2$ $\gamma = 0$ dB
4B 25% most important data rate $4-2/L$	Fig. 7 $P = 3.07,\ PAR = 1{\cdot}7$ $d^2(\Omega_{ab},\Omega_{cd}) \geq \begin{cases} 4, & \text{if } ab = \bar{c}d \\ 3, & \text{if } ab = \bar{c}\bar{d} \\ 1, & \text{if } ab = c\bar{d} \end{cases}$	[6,31]: 16-state, rate 1/2 convolutional code $d^2(C_0) = 14$ $\Gamma = 7.6$ dB	single parity check code over FF_4 with redundancy $2/L$ $d^2(C_1) = 4$ $\gamma = 2.12$ dB
4C 25% most important data rate $4-1/L$	Fig. 11 $P = 5.25,\ PAR = 1{\cdot}7$ $d^2(\Omega_{ab},\Omega_{cd}) \geq \begin{cases} 10, & \text{if } ab = \bar{c}\bar{d} \\ 2, & \text{if } ab \neq cd \end{cases}$	[23,35]: see Example 4A $d^2(C_0) = 26$ $\Gamma = 7.91$ dB	single parity check code with redundancy $1/L$ $d^2(C_1) = 4$ $\gamma = 0.21$ dB
4D 25% most important data rate $4-2/L$	Fig. 12 $P = 11,\ PAR = 1{\cdot}86$ $d^2(\Omega_0,\Omega_1) = 25$	single parity check code with redundancy $1/L$ $d^2(C_0) = 50$ $\Gamma = 7.54$ dB	[23,35]: see Example 4A & parity check $d^2(C_0) = 14$ $\gamma = 2.20$ dB
4E 25% most important data rate $4-1/L$	Fig. 9 $P = (4+24x+140x^2)/32$ $d^2(\Omega_{ab},\Omega_{cd}) \geq \begin{cases} 1+x^2, & \text{if } ab = \bar{c}\bar{d} \\ x^2, & \text{if } ab = c\bar{d} \\ 1, & \text{if } ab = \bar{c}d \end{cases}$ $PAR = 2{\cdot}05$ for $x = 0.3$	$\underline{x}$ $\underline{\text{code}}$ $\underline{\Gamma(\text{dB})}$ 0.4 [23,35] 4.92 0.3 [23,35] 6.11 0.2 [23,35] 7.90 0.1 [23,35] 10.10	2-level code: see Example 4D $d^2(C_0) = 14x^2$ $\underline{x}$ $\underline{\gamma(\text{dB})}$ 0.4 3.96 0.3 3.25 0.2 1.91 0.10 −1.46

TABLE 2

Example	Signal Constellation	Code C_0 for important bits	Code C_1 for less important bits
4F 50% important data rate $4-1/L$	Fig. 13 $P = 10.5$, $PAR = 2{\cdot}33$ $d^2(\Omega_{ab}, \Omega_{c}) \geq \begin{cases} 5, \text{ if } ab = c\bar{d} \\ 2, \text{ if } ab = \bar{c}\bar{d} \\ 1, \text{ if } ab = \bar{c}d \end{cases}$	[2,31]: 16-state, rate 1/2 convolutional code $d^2(C_0) = 16$ $\Gamma = 5.8$ dB	single parity check code with redundancy $1/L$ $d^2(C_1) = 8$ $\gamma = 2.8$ dB
4G 33% important data rate $4-2/L$	Fig. 11 $P = 5.25$, $PAR = 1{\cdot}7$ see Example 4C	2-level code: - [52,66,76]: 16-state rate 1/3 convolutional code, $d_H = 12$ & parity check $d^2(C_0) = \min(12\times2, 2\times10)$ $= 20$ $\Gamma = 6.8$ dB	2-level code: - [31,33] with puncturing rule $\begin{bmatrix} 1 & 1 \\ 1 & 0 \end{bmatrix}$, to give 16-state, rate 2/3 code, $d_H = 5$ & parity check $d^2(C_1) =$ $\min(5\times8, 2\times16, 64) = 32$ $\gamma = 2.8$ dB
4H 37.5% important data rate $4-1/L$	Fig. 10 $P = 5.25$, $PAR = 1{\cdot}76$ see Example 4C	2-level code: - [23,35]: 16-state rate 1/2 convolutional code, $d_H = 7$ & parity check $d^2(C_0) = \min(7\times2, 2\times10)$ $= 14$ $\Gamma = 5.3$ dB	2-level code: - rate 2/3 punctured convolutional code from Example 4G & 16-state, rate 7/8 punctured convolutional code with $d_H = 3$ $d^2(C_1) =$ $\min(40, 3\times16, 64) = 40$ $\gamma = 3.76$ dB

TABLE 2 Cont'd.

Example	Signal Constellation	Code C_0 for important bits	Code C_1 for less important bits
4J 43.75% important data rate $4+1/4-2/L$	Fig. 4 $P = 4.75$, $PAR = 1{\cdot}9$ $d^2(\Omega_{ab}, \Omega_{cd}) \geq \begin{cases} 8, & \text{if } ab = \overline{cd} \\ 4, & \text{if } ab \neq cd \end{cases}$	2-level code: - rate 3/4 convolutional code from Example 4F & parity check $d^2(C_0) = \min(4\times4, 2\times8) = 16$ $\Gamma = 6.24$ dB	3-level code: - rate 1/2 convolutional code from Example 4A & rate 3/4 convolutional code from Example 4F & parity check $d^2(C_1) = 7$ $\gamma = 2.7$ dB
4K 56.25% important data rate $4-2/L$	Fig. 14 $P = 16\frac{23}{32}$, $PAR = 1{\cdot}63$ $d^2(\Omega_{abc}, \Omega_{def}) \geq \begin{cases} 9, & \text{if } c = \bar{f} \\ 18, & \text{if } bc = \overline{ef} \\ 36, & \text{if } abc = \overline{def} \end{cases}$	3-level code: - rate 1/2 convolutional code from Example 4A & rate 3/4 convolutional code from Example 4F & parity check $d^2(C_0) = \min(63, 72, 72) = 63$ $\Gamma = 6.73$ dB	2-level code: - rate 3/4 convolutional code from Example 4F & parity check $d^2(C_1) = 16$ $\gamma = 0.78$ dB
4L 22.5% important data rate $3+7/8-1/L$ (power penalty) 0.75 dB	Fig. 12 $P = 11$, $PAR = 1{\cdot}86$ $d^2(\Omega_0, \Omega_1) = 25$	[23,35] with puncturing rule [1010011, 110010] to give 16-state rate 7/8 convolutional code with $d_H = 3$ $d^2(C_0) = 75$ $\Gamma = 8.55$ dB	see Example 4D $d^2(C_1) = 14$ $\gamma = 1.27$ dB

In these tables, P denotes the average power per dimension; PAR denotes the peak-to-average power ratio; d^2 denotes the square Euclidian distance between the supersymbols represented by its arguments; Γ denotes the nominal coding gain for the more important bits; γ denotes the nominal coding gain for the less important bits; and $[g_1, g_2]$ is the generator matrix, in octal notation, for the convolutional codes indicated; L is the length of the parity check code. For non-uniform constellations, the degree of non-uniformity is determined by x. The tables also show the achieveble gains as a function of x.

In designing unequal error protection signaling schemes, such as those just presented in TABLES 1 and 2, one typically must first be given the values of certain paremeters. These include (a) available channel bandwidth, (b) the worst-case channel SNR, (c) the number of classes of bits, (d) the percentage of bits in each class, (e) the desired quality of the final received signal under the worst-case channel conditions, (f) acceptable

decoder complexity, and (g) peek-to-average power ratio. One can then proceed to design an unequal-error-protection scheme consistent with the given parameter values.

Typically, one might begin by choosing the signal constellation to allow for about one overall bit of redundancy per symbol. The number of supersymbols that is needed is determined by the actual number of important bits per symbol. For example, if more than 25%, and less than 50%, of the bits are important, and if the available bandwidth and required overall bitrate dictate, say, four information bits per symbol, then we will have to transmit one important bit per symbol. Assuming, further, that it is desired to provide some amount of redundancy for the important bits and that those bits will be used to select the supersymbols, then it may be reasonable to use a two-dimensional constellation having four supersymbols. Various of the constellations disclosed herein, or any other desired constellation, can be used as the initial design choice. One would then proceed to select codings to be used for the bit streams depending on the relative redundancies desired to be allocated to them Importantly, the use of multi-level codes, in accordance with the present invention, can facilitate the attainment of the desired design criteria. If, upon analysis, it appears thet the desired quality at the worst-case SNR is not achieved with the design arrived at to this point, other constellations and other codes can be explored with a view toward seeing whether it can, in fact, be achieved. If this does not appear possible, one or more of the previously given constraints, such as acceptable decoder complexity, may have to be relaxed. Other design criteria are conceivable as well as other parameters that may affect the choice of the code design.

In accordance with an advantageous technique for assigning the values of the bits that are used to select the symbols from each identified supersymbol to those symbols, a pair of symbols from respective different supersymbols are assigned to the same less-important-bit values if the distance between that pair of symbols is the minimum distance between any pair of symbols of those two supersymbols. Such a scheme is illustrated in FIG. 10, in which, like in FIG. 4, each symbol is lebelled with a b_2, b_1, b_0 set of values. Note, for example, that the symbols labelled 110 in eech of the supersymbols Ω_{00}, Ω_{01}, Ω_{10} and Ω_{11} all setisfy the aforementioned criterion. Such a labelling scheme--which is achievable to varying degrees, depending on the constellation and supersymbol geometries--is adventegeous in that it improves the probability that the less-important-bit values will be decoded properly, even if an error is made in recovering the correct supersymbol sequence. Beyond this, it may be possible to achieve former benefit by similar judicious choice of bit assignments for those symbols which do not meet the above-mentioned minimum distence criterion. This would need to be done, however, within the constraints that are imposed by any coding thet is implemented for the symbols within the supersymbol, as was, in fact, the case of the example described above in conjunction with the constellation of FIG. 4.

The foregoing merely illustrates the principles of the present invention. For example, although the illustrative embodiments are implemented using two data streams--the most- end less-important--the invention can be used in schemes which include three or more streams. Additionally, although two-dimensional constellations ere shown, the invention can be used in schemes using constellations with more than two dimensions.

It may elso be noted that although the invention is illustrated herein es being implemented with discrete functional building blocks, e.g., source coders, scramblers, etc., the functions of any one or more of those building blocks can be carried out using one or more appropriate programmed processors, digital signal processing (DSP) chips, etc.

Claims

1. Apparatus for communicating information comprising
   means (104) for generating a digital signal representing the information, the digital signal being comprised of at least first and second streams of data elements,
   CHARACTERIZED BY
   means (110, 111, 114, 115, 131) for chennel mapping the digital signal in such a way that the probability of channel-induced error for the data elements of the first data stream is less than the probability of channel-induced error for the data elements of the second data stream, said channel mapping means including means (114) for multi-level coding at least one of said streams, and
   means (141, 151) for transmitting the channel mapped signal over a communication channel.

2. The invention of claim 1 wherein said information is television signal information and wherein first stream of data represents television signal information that is more important than the television signal information represented by the data elements of said second stream.

3. The invention of claim 1 wherein said et least one of said streams includes at least two substreams, and

wherein said multi-level coding means includes means (1141, 1142) for redundancy coding at least one of said substreams and for combining all of the redundancy coded substreams of said one stream and any of its substreams that are not redundancy coded to form a coded signal for use in the channel mapping.

4. The invention of claim 3 wherein the channel mapping means selects a sequence of symbols from a predefined constellation to represent the data elements, the constellation being comprised of supersymbols, the minimum distance between the symbols within at least one of the supersymbols being less than the maximum intra-subset distance for said constellation.

5. The invention of claim 4 wherein the minimum distance between the symbols of said at least one of the supersymbols is greater than the minimum distance between symbols of the constellation as a whole.

FIG. 1

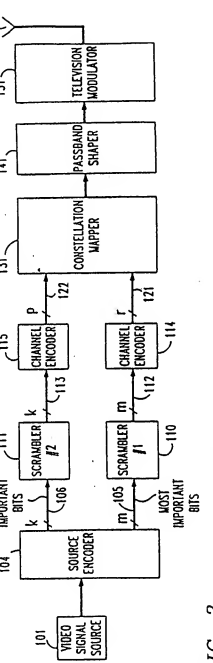

101
VIDEO SIGNAL SOURCE
104
SOURCE ENCODER
k
m
LESS IMPORTANT BITS
106
105
MOST IMPORTANT BITS
111
SCRAMBLER #2
110
SCRAMBLER #1
k
113
m
112
115
CHANNEL ENCODER
114
CHANNEL ENCODER
p
122
r
121
131
CONSTELLATION MAPPER
141
PASSBAND SHAPER
151
TELEVISION MODULATOR
152

FIG. 2

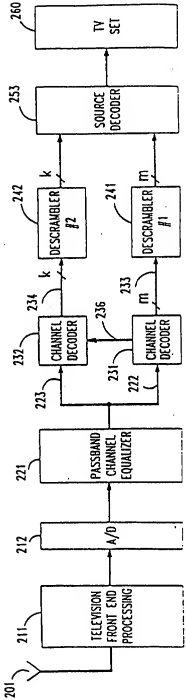

201
211
TELEVISION FRONT END PROCESSING
212
A/D
221
PASSBAND CHANNEL EQUALIZER
223
222
232
CHANNEL DECODER
231
CHANNEL DECODER
236
234
k
233
m
242
DESCRAMBLER #2
241
DESCRAMBLER #1
k
m
253
SOURCE DECODER
260
TV SET

FIG. 3

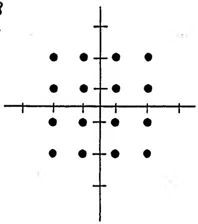


FIG. 4

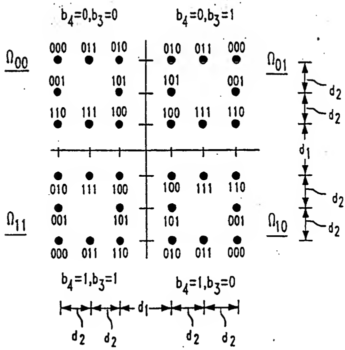

$b_4=0, b_3=0$
$b_4=0, b_3=1$
Ω_{00}
Ω_{01}
000 011 010
010 011 000
001 101
101 001
110 111 100
100 111 110
010 111 100
100 111 110
001 101
101 001
000 011 110
010 011 000
Ω_{11}
Ω_{10}
$b_4=1, b_3=1$
$b_4=1, b_3=0$
d_1
d_2

FIG. 5

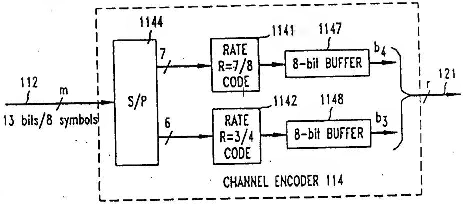

1144
1141
1147
7
RATE R=7/8 CODE
8-bit BUFFER
b_4
112
m
13 bits/8 symbols
S/P
121
r
1142
1148
6
RATE R=3/4 CODE
8-bit BUFFER
b_3
CHANNEL ENCODER 114

FIG. 6

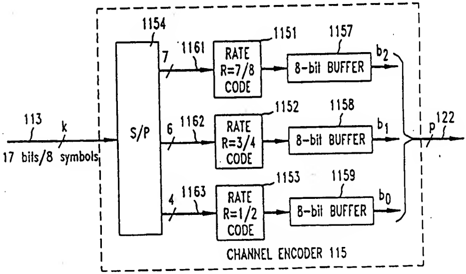

1154
1151
1157
7
1161
RATE R=7/8 CODE
8-bit BUFFER
b_2
1152
1158
113
k
17 bits/8 symbols
S/P
6
1162
RATE R=3/4 CODE
8-bit BUFFER
b_1
p
122
1153
1159
4
1163
RATE R=1/2 CODE
8-bit BUFFER
b_0
CHANNEL ENCODER 115

FIG. 7

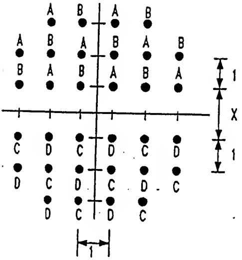

A
B
A
B
A
B
A
B
A
B
B
A
B
A
B
A
1
X
1
C
D
C
D
C
D
D
C
D
C
D
C
D
C
D
C
1

FIG. 8

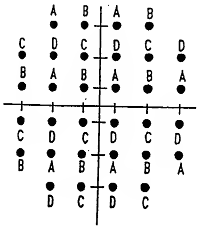
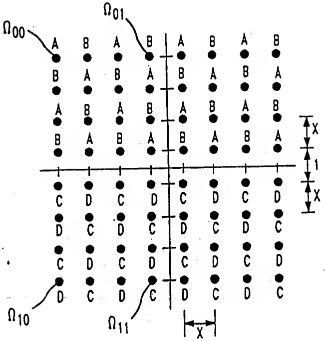
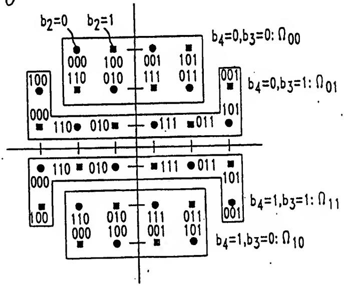

A
B
A
B
C
D
C
D
C
D
B
A
B
A
B
A
C
D
C
D
C
D
B
A
B
A
B
A
D
C
D
C

FIG. 9



Ω00
Ω01
A B A B A B A B
B A B A B A B A
A B A B A B A B
B A B A B A B A
X
1
X
C D C D C D C D
D C D C D C D C
C D C D C D C D
D C D C D C D C
Ω10
Ω11
X

FIG. 10



b2=0
b2=1
b4=0,b3=0: Ω00
000 100 001 101
110 010 111 011
b4=0,b3=1: Ω01
100
000
110 010 111 011
001
101
110 010 111 011
000
100
101
001
b4=1,b3=1: Ω11
110 010 111 011
000 100 001 101
b4=1,b3=0: Ω10

FIG. 11

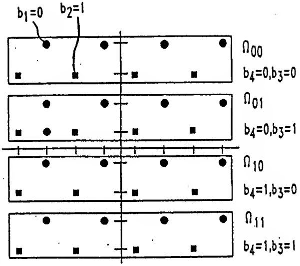

$b_1=0$
$b_2=1$
Ω_{00}
$b_4=0,b_3=0$
Ω_{01}
$b_4=0,b_3=1$
Ω_{10}
$b_4=1,b_3=0$
Ω_{11}
$b_4=1,b_3=1$

FIG. 12

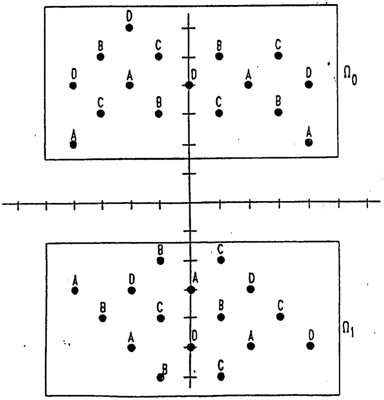

Ω_0
Ω_1

FIG. 13

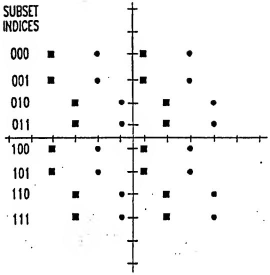

SUBSET
INDICES
000
001
010
011
100
101
110
111

FIG. 14

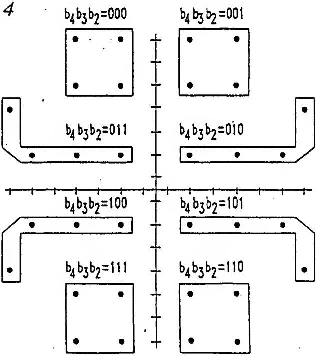

$b_4b_3b_2$=000
$b_4b_3b_2$=001
$b_4b_3b_2$=011
$b_4b_3b_2$=010
$b_4b_3b_2$=100
$b_4b_3b_2$=101
$b_4b_3b_2$=111
$b_4b_3b_2$=110

FIG. 15

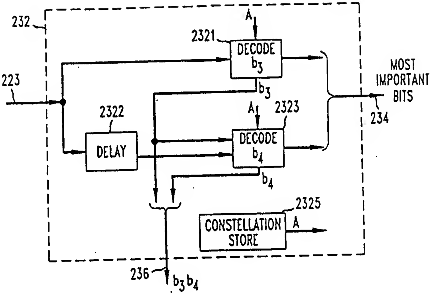
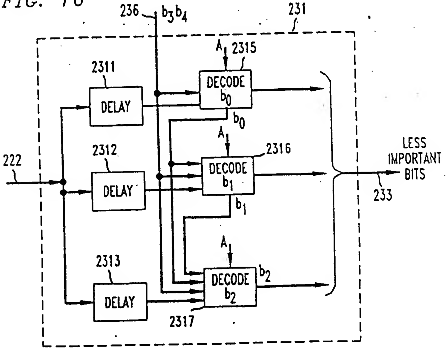

232
223
2321
A
DECODE
b_3
b_3
2322
DELAY
A
2323
DECODE
b_4
b_4
2325
CONSTELLATION
STORE
A
MOST
IMPORTANT
BITS
234
236
b_3b_4

FIG. 16


236
b_3b_4
231
A
2315
2311
DELAY
DECODE
b_0
b_0
A
2312
2316
222
DELAY
DECODE
b_1
b_1
A
2313
DELAY
DECODE
b_2
b_2
2317
LESS
IMPORTANT
BITS
233

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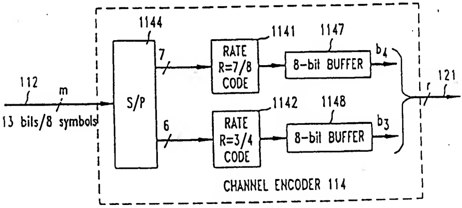
(54) Coded modulation with unequal error protection.

(57) Digital signals (from 101), such as digital television signals, are subjected to a source coding step (by 104) followed by a channel mapping step (by 114 and 115). The source coding step causes the signal to be represented by first and second data streams (on 105, 106). The first stream carries data regarded as more important and the second carries data regarded as less important. In the channel mapping step, the mapping is such that the data elements of the various data streams have differing probabilities of being erroneously detected at the receiver. The channel mapping step includes at least one multi-level coding step. The signal constellations used in the channel mapping step are partitioned into supersymbols, in which the distance between the symbols comprising at least ones of the supersymbols is less than a parameter referred to as the maximum intra-subset distance (MID). Additionally, in some constellations, the minimum distance between at least ones of the symbols of at least one of the supersymbols is greater than the minimum distance between the symbols of the constellation as a whole while, again, still being less than the MID. The first data stream is used to identify a sequence of supersymbols, while the second data stream is used to select particular symbols from the identified supersymbols.

FIG. 5



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Application Number

EP 92 30 9608
Page 1

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Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.5)
E	EP-A-0 540 232 (AT & T) * page 3, line 12 - line 20 * * page 5, line 20 - line 29 * * page 6, line 29 - line 36 * * page 7, line 23 - line 32 * * page 7, line 43 - line 49 * * page 8, line 50 - line 58 * * figures 2,5,6,15,16 * ---	1-5	H04L27/34
X	Proceedings of the Fourth International Workshop on HDTV and Beyond, 04.-06.09.1991, Turin, IT, SIGNAL PROCESSING of HDTV, III, pages 61-69, Elsevier, Amsterdam, NL; K. M. UZ et al.: 'Multiresolution Source and Channel Coding for Digital Broadcast of HDTV.' * figure 2 * * page 63, paragraph 4 - page 65, paragraph 1 * ---	1-5	
			TECHNICAL FIELDS SEARCHED (Int. Cl.5)
P,X	SMPTE JOURNAL vol. 101, no. 8, August 1992, SCARSDALE, NY US pages 538 - 549 W. F. SCHREIBER: 'Spread-Spectrum Television Broadcasting.' * figure 3 * * page 539, middle column, line 11 - line 36 * * page 539, middle column, line 46 - line 49 * * page 542, left column, paragraph 2 * * page 543, right column, paragraph 2 * --- -/--	1-5	H04L ,

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	15 JUNE 1993	GRIES T.M.

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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A : technological background
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European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 92 30 9608
Page 2

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P,X	SIGNAL PROCESSING. IMAGE COMMUNICATION vol. 4, August 1992, AMSTERDAM NL pages 283 - 292 K. M. UZ ET AL.: 'Combined multiresolution source coding and modulation for digital broadcast of HDTV.' * figure 2 * * page 285, left column, paragraph 3 - page 287, line 2 * ---	1-5	
P,X	MULTIDIMENSIONAL SYSTEMS AND SIGNAL PROCESSING, vol. 3, no. 2-3, May 1992, NL, pages 161-187; M. VETTERLI / K. M. UZ: 'Multiresolution Coding Techniques for Digital Television: A Review.' * page 182-183, section 6.3: 'Multiresolution transmission' * * figure 14 * ---	1-5	
P,X	EP-A-0 485 108 (AT & T) * column 1, line 6 - column 3, line 33 * * column 6, line 21 - column 8, line 40 * * column 13, line 28 - line 45 * * column 14, line 4 - line 9 * * claims 1,2,4,5,7-10,12-15,17; figures 1,2,4,5,8,12 * ---	1-5	TECHNICAL FIELDS SEARCHED (Int. Cl.5)
P,X	EP-A-0 485 105 (AT & T) * page 2, line 32 - page 3, line 6 * * page 4, line 6 - line 38 * * figures 5,6,8-11 * --- -/--	1-5	

The present search report has been drawn up for all claims

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EUROPEAN SEARCH REPORT

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EP 92 30 9608
Page 3

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.5)
A	EP-A-0 282 298 (FORD AEROSPACE & COMMUNICATIONS CORPORATION) * page 3, line 14 - line 41 * * page 5, line 30 - line 37 * * page 5, line 55 - line 60 * * page 6, line 50 - line 63 * * page 7, line 32 - page 8, line 39 * * figures 5-7,13,14 * ---	1,3-5	
D,A	IEEE TRANSACTIONS ON COMMUNICATIONS vol. 37, no. 3, March 1989, NEW YORK US pages 222 - 229 A. R. CALDERBANK: 'Multilevel Codes and Multistage Decoding.' * page 222, left column, paragraph 5 - right column, paragraph 2 * * page 222, right column, paragraph 5 - page 223, left column, line 3 * * page 223, left column, paragraph 7 * ---	1,3	
A	IEEE TRANSACTIONS ON INFORMATION THEORY vol. 23, no. 3, 1977, NEW YORK US pages 371 - 377 H. IMAI / S. HIRAKAWA: 'A New Multilevel Coding Method Using Error-Correcting Codes.' * figures * * page 371, right column, paragraph 3 - page 372, left column, paragraph 3 * -----	1,3	TECHNICAL FIELDS SEARCHED (Int. Cl.5)

The present search report has been drawn up for all claims

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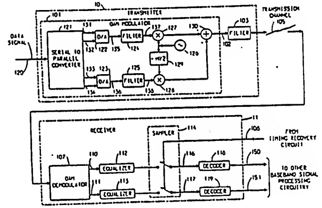
(54) Title: SINGLE-SIDEBAND COMMUNICATION SYSTEM



27/40

(57) Abstract

A bandwidth reduction technique for use in digital systems wherein elements of a data signal modulate quadrature-related carriers. This modulation, referred to as quadrature amplitude modulation (QAM) or phase shift keying (PSK), generates a double-sideband signal which is transmitted in a variety of communications systems. In accordance with the present invention, the above-described double-sideband signal is filtered (103) to form a single-sideband signal prior to transmission. While this use of a single-sideband signal, in lieu of a double-sideband signal, effectively doubles the system capacity by permitting the use of two communications systems in the bandwidth previously occupied by one system, the filtering process (103) contaminates the data signal elements. To recover the data signal elements at the receiver, received signal elements are formed by extracting the carrier signals (107). Next, these received signal elements are altered by preselected quantities to form estimates of each data signal element (301, 302...307). A comparison of the formed estimates (318) against the set of permissible values for each data signal element then determines which estimate is correct.

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SINGLE-SIDEBAND COMMUNICATION SYSTEM

Technical Field

The present invention relates to a digital communications system which transmits a single-sideband signal comprising modulated quadrature-related carriers.

Background of the Invention

Digital communication systems utilize a myriad of modulation formats. In one commonly-used format, elements of a data signal modulate quadrature-related carrier signals. This type of modulation has a variety of names, such as phase shift keying (PSK), quadrature amplitude modulation (QAM), and asynchronous phase shift keying (APSK). The information conveyed by the data signal is, of course, virtually limitless and can include voice, video, facsimile and the like. Moreover, the transmission channel carrying the modulated carriers is also not limited and, at present, may include air, wire or lightguide.

A problem in practically all communications systems is that the transmission channel is band-limited. That is, there is a finite frequency interval which can be used to convey information. This limitation arises because of system and/or device requirements. While the severity of this problem does vary from system to system, it still can be said that the ability to convey still more information in a given frequency interval would be highly desirable.

One technique to increase the information-carrying capacity of a digital system transmitting modulated quadrature-related carriers is to increase the number of permissible modulation states. An example of this technique is exemplified by the design and deployment of 64 QAM systems in lieu of 16 QAM systems in applications requiring greater capacity. The problem with this technique is that the change in the number of modulation states requires at least the design and development of new modulators and demodulators. This effort is often

expensive and the resulting equipment, at times, can not be retrofitted into operational systems without great expense.

Another technique to increase system capacity has been to utilize single-sideband signals instead of double-sideband signals. This technique is rather simple to implement and has been routinely used in formats which modulate a single carrier signal. Unfortunately, this technique has not been used for systems utilizing quadrature-related carriers because there was no known way of intelligently decoding the received signal after single-sidebanding.

Summary of the Invention

The present invention is intended for use in digital communications systems wherein elements of a data signal modulate quadrature-related carrier signals. To reduce the required bandwidth, the resulting modulated quadrature-related carriers are transformed into a single-sideband signal. After propagation through the transmission channel, the received single-sideband signal is demodulated into received signal elements. Each of these elements includes an element of the data signal along with a spurious signal introduced by the single-sideband transformation. To recover the data signal elements, each received signal element is altered to form at least one estimate of the corresponding data signal element. Each estimate formed is then compared against a set of permissible data signal element values and the estimate is outputted if a preselected criterion is met.

A feature of the present invention is that it can be implemented within existing digital communications systems to provide a substantial increase in information-carrying capacity within some preselected bandwidth.

A further feature of the present invention is that it can be used with conventional demodulation and equalization techniques.

Brief Description of the Drawing

FIG. 1 is a block schematic diagram of a

communications system which incorporates the present invention;

FIG. 2 is a plot of the signal space diagram of the signal levels transmitted by the communications system of FIG. 1; and

FIG. 3 is a detailed schematic diagram of decoders 118 or 119 shown in the communications system of FIG. 1.

Detailed Description

FIG. 1 shows an exemplary QAM communications system which incorporates the present invention. At transmitter 10, a digital data signal on lead 120 is coupled to QAM modulator 101. Within modulator 101, serial-to-parallel converter 121 spreads successive data signals on lead 120 over four paths 131,132,133, and 134. Digital-to-analog (D/A) converter 122 quantizes the signals appearing on leads 131 and 132 into a number of signal voltages which appear on lead 135. Similarly, D/A converter 132 quantizes the signals on leads 133 and 134 into a number of signal voltages which are coupled to lead 136. Multipliers 127 and 128 receive the signal voltages on leads 135 and 136 after they are respectively smoothed by Nyquist filters 124 and 125. Multiplier 127 modulates the amplitude of a carrier signal generated by oscillator 126 with the signals on lead 135 after filtering. In similar fashion, multiplier 128 modulates the amplitude of a second carrier signal with the signals on lead 136 after smoothing by Nyquist filter 125. The second carrier signal supplied to multiplier 128 is generated by shifting the carrier signal generated by oscillator 126 by minus $\pi/2$ radians via phase shifter 129. Hence, the pair of carrier signals supplied to multipliers 127 and 128 are in spatial quadrature to one another and the products provided by multipliers 128 and 129 are each double-sideband signals. Summer 130 then adds the products provided by multipliers 128 and 129 and outputs this sum, also a double-sideband signal onto

lead 102.

Reviewing the signal processing provided by the transmitter components discussed thus far, it can be said that these components modulate quadrature-related carriers with elements of a data signal, wherein one element of the data signal comprises the signals appearing on leads 131,132 or 135 or 137 while the other data signal element comprises the signals appearing on leads 133,134 or 136 or 138. In addition, if we select the number and permitted values of the signal voltages provided by D/A converters 122 and 123, we can graphically depict all of the possible combinations of transmitted carrier signal amplitudes which represent the data signal on a cartesian coordinate plot. Such a plot is commonly referred to as a signal space diagram.

Refer now to FIG. 2 which shows the signal space diagram for the illustrative transmitter of FIG. 1. The data signal element appearing on lead 137 is designated as the "I" or in-phase element of the data signal while the data signal element appearing on lead 138 is referred to as the "Q" or quadrature element. As shown, the permissible values of the "I" and "Q" elements are ± 1 and ± 3 volts and all possible combinations of these permissible values form 16 signal states, designated as 201, in FIG. 2.

In prior art communications systems, the output of summer 130 is coupled to a transmission channel which propagates the information to system receiver 11. In accordance with the present invention, a filter 103 is added to the transmitter to convert the double-sideband signal at the output of summer 130 into a single-sideband signal thereby reducing the bandwidth required for signal transmission. This bandwidth reduction also permits the transmission of a second single-sideband QAM signal in the recovered frequency interval. The resulting capacity of two 16 QAM single-sideband signals is equivalent to that of a 256 QAM double-sideband signal. The double-sideband to single-sideband signal conversion, however, corrupts the

operation of conventional QAM receiver circuitry and additional functional capability is required in the receiver to intelligently recover the data signal elements. At this juncture, it should be understood that the present invention is also applicable to radio systems wherein additional circuitry is often disposed between summer 130 and the transmission channel to shift the frequency of the transmitted carriers to a higher frequency band. Moreover, the present invention is not limited to QAM systems and, indeed, may be utilized in any system which transmits a signal comprising modulated quadrature-related carriers which are modulated in phase or amplitude or some combination of phase and amplitude.

To understand the principles of the present invention, it is first necessary to consider the effects of filtering one of the sidebands of the illustrative double-sideband QAM signal and then transmitting the resulting single-sideband signal through a transmission channel.

The QAM signal appearing at the output of summer 130 can be expressed as a function of time s(t) with

$$s(t) = i(t) \cos w_c t - q(t) \sin w_c t \; ; \qquad (1)$$

and where w_c denotes the frequency of the carrier generated by oscillator 126, and

$i(t)$ and $q(t)$ respectively denote the values of the I and Q data signal elements as a function of time.

When s(t) is passed through filter 103 with an impulse response h(t) in order to reject either one of the sidebands, we can express the resulting single-sideband signal as $[s(t)]_{SSB}$ with

$$[s(t)]_{SSB} = \int_{-\infty}^{+\infty} h(\tau) i(t-\tau) \cos[w_c(t-\tau)] d\tau - \int_{-\infty}^{+\infty} h(\tau) q(t-\tau) \sin[w_c(t-\tau)] d\tau \qquad (2)$$

and where τ represents a dummy variable of integration.

Using the trigonometric identities:

$$\cos[w_c(t-\tau)] = \cos w_c t \cos w_c \tau + \sin w_c t \sin w_c \tau \text{ and}$$
$$\sin[w_c(t-\tau)] = \sin w_c t \cos w_c \tau - \cos w_c t \sin w_c \tau, \quad (3)$$

equation (2) can be rewritten as:

$$\begin{aligned}[s(t)]_{SSB} &= \int_{-\infty}^{+\infty}\{h(\tau)i(t-\tau)\cos w_c\tau \, d\tau\}\cos w_c t \\ &+ \int_{-\infty}^{+\infty}\{h(\tau)q(t-\tau)\sin w_c\tau \, d\tau\}\cos w_c t \\ &+ \int_{-\infty}^{+\infty}\{h(\tau)i(t-\tau)\sin w_c\tau \, d\tau\}\sin w_c t \\ &- \int_{-\infty}^{+\infty}\{h(\tau)q(t-\tau)\cos w_c\tau \, d\tau\}\sin w_c t \qquad (4)\end{aligned}$$

Equation (4), in turn, can be written as:

$$\begin{aligned}[s(t)]_{SSB} = \tfrac{1}{2}\ \{i(t) + \hat{q}(t)\}\cos w_c t \\ - \tfrac{1}{2}\{q(t) - \hat{i}(t)\}\sin w_c t\ , \qquad (5)\end{aligned}$$

where $\hat{i}(t)$ and $\hat{q}(t)$ are the Hilbert transforms of $i(t)$ and $q(t)$, respectively.

A comparison of equation (5) with equation (1) reveals that the effect of eliminating one of the sidebands of the QAM signal of equation (1) contaminates $i(t)$ with the Hilbert transform of $q(t)$ and contaminates $q(t)$ with the Hilbert transform of $i(t)$. Consequently, the receiver of FIG. 1 must be provided with the capability of eliminating $\hat{q}(t)$ and $\hat{i}(t)$ to respectively recover the $i(t)$ and $q(t)$ components.

Refer back to FIG. 1 and consider the general case where transmission channel 105 is dispersive and introduces distortion comprising intersymbol interference (ISI), cross-rail interference (X-rail ISI) and Gaussian noise ($n(t)$). If $s(t)_{SSB}$ is coupled through a conventional QAM demodulator 107, two received data elements $i'(t)$ and $q'(t)$ are formed on leads 110 and 111. The generation of $i'(t)$ and $q'(t)$ is accomplished by extracting the quadrature-related carriers from the received signal using well-known carrier recovery

techniques. The signals on leads 110 and 111 can be expressed as:

$i'(t) = [i(t) + \hat{q}(t)] + ISI + X\text{-rail } ISI + n_I(t)$, (6)

and

$q'(t) = [q(t) - \hat{i}(t)] + ISI + X\text{-rail } ISI + n_Q(t)$, (7)

with $n_I(t)$ and $n_Q(t)$ respectively representing the Gaussian noise introduced into i(t) and q(t).

The ISI and X-rail ISI in equations (6) and (7) can be eliminated by coupling i'(t) and q'(t) through conventional transversal equalizers 112 and 113 which are configured to operate on i'(t) and q'(t) as if $[i(t) + \hat{q}(t)]$ and $[q(t) - \hat{i}(t)]$ were the information signals. The equalized signals $i_E(t)$ and $q_E(t)$ appearing at the output of equalizers 112 and 113 are then sampled at the baud rate, 1/T, by sampler 114. The k^{th} sample, where K is any integer, can be expressed as

$i_E(kT) = [i(kT) + \hat{q}(kT)] + n_{IE}(kT)$ (8)

for lead 116 and

$q_E(kT) = q(kT) - \hat{i}(kT) + n_{QE}(kT)$ (9)

for lead 117. The expressions $n_{IE}(kT)$ and $n_{QE}(kT)$ represent the Gaussian noise in the received signal components after equalization. Sampler 114 is controlled by a timing signal on lead 108 which is supplied by conventional timing recovery circuitry (not shown) in the receiver.

To recover the information carrying components of i(kT) and q(kT), $\hat{q}(kT)$ and $\hat{i}(kT)$ must be eliminated. It can be shown that $\hat{q}(kT)$ and $\hat{i}(kT)$ can only assume a limited number of values and the values are a function of the quantized values provided by D/A converters 122 and 123. The set of values for $\hat{i}(kT)$ and $\hat{q}(kT)$ for any communications system utilizing Nyquist filtering can be expressed as

$\hat{i}(kT) = -1/2q((k-1)T) + 1/2q((k+1)T)$ (10)

and

$$\hat{q}(kT) = -1/2i((k-1)T) + 1/2i((k+1)T). \tag{11}$$

That is, the Hilbert transform of i(t) at the k^{th} sampling time is a function of q(t) at the (k-1) and (k+1) sampling times wherein the (k-1) and (k+1) sampling times are respectively one sampling time immediately preceding and one sampling time immediately succeeding the k^{th} sampling time. And, the Hilbert transform of q(t) at the k^{th} sampling time is a function of i(t) at the (k-1) and (k+1) sampling times wherein the (k-1) and (k+1) sampling times are respectively one sampling time immediately preceding and one sampling time immediately succeeding the k^{th} sampling time.

From equations (10) and (11), it follows that in the illustrative 16 QAM communication system wherein i(t) and q(t) can take on the values of $\pm$ 1 and $\pm$ 3 volts, $\hat{i}(kT)$ and $\hat{q}(kT)$ can take on any value from the set {0, -1, -2, -3, 1, 2, 3}. Therefore, at any sampling instant, kT, $\hat{i}(kT)$ and $\hat{q}(kT)$ can assume one of seven possible values.

Refer now to FIG. 3 which shows a detailed schematic of the circuitry within decoders 118 and 119 of FIG. 1. In decoder 118, the k^{th} sample $i_E(kT)$ is supplied to seven summers 301, 302, ... 307 to form seven estimates of i(kT) on leads 311 through 317. Each summer forms one of these estimates by subtracting a different one of the seven possible values of $\hat{q}(t)$ from $i_E(kT)$. Each of leads 321-327 supplies a different value of $\hat{q}(t)$ from a source of reference voltages (not shown). Selection circuit 318, comprising multiple threshold detectors, compares each estimate against the permissible values of i(t), namely, $\pm$ 1 and $\pm$ 3 volts, and selects the estimate of i(kT) which is closest to any of the permissible values. This selected estimate is outputted on lead 150. Decoder 119 performs an identical operation on each sample $q_E(kT)$, with the estimate of q(kT) closest to one of the permissible values of q(t) being outputted on

lead 151 in FIG. 1. As shown, leads 150 and 151 couples the selected estimates of i(t) and q(t) to timing recovery and other receiver circuitry for further signal processing not connected with the present invention.

In the process of estimate formation and selection, it is possible for ambiguities to arise, i.e., there are two or more estimates formed which are equally close to different permissible data element values. This problem can be avoided by using one set of values for i(t) and a different set of values for q(t). For example, for the illustrative 16 QAM signal constellation shown in FIG. 2, values of i(t) equal to ± 1 and ± 3 volts and the values of q(t) equal to ± 1.5 and ± 4.5 volts provide signal states 201' which circumvent the aforesaid ambiguity problem.

While the disclosed decoders 118 and 119 comprise circuitry which simultaneously provides seven possible estimates of i(t) and q(t) using parallel signal processing, the decoders could comprise only one adder which sequentially forms seven estimates of i(t) or q(t). In this approach, selection circuit 318 compares each estimate against the permissible values of a data element and any estimate which falls within a predetermined interval surrounding each permissible value would be outputted. Upon selecting an estimate, selector circuit 318 would inhibit the outputting of any other estimate until the next sample is received from sampler 114.

It should, of course, be understood that the present invention is not limited to the particular embodiment disclosed and that numerous modifications will occur to those skilled in the art which are within the spirit and scope of the invention. First, for example, the use of transversal equalizers in the receiver is not required if the magnitude of ISI and X-rail ISI is not large relative to the difference between permissible data element values. This is often true in lightwave and wire

systems wherein the transfer function of the transmission channel is not time-varying. Second, while Nyquist filters are only shown in transmitter 10, half-Nyquist filters could also be utilized in transmitter 10 and receiver 11.

Claims

1. Receiver apparatus for use in a digital transmission system wherein elements of a data signal modulate quadrature-related carriers and wherein said carriers are transformed into a single-sideband signal, said receiver apparatus comprising

means for demodulating said single-sideband signal to form received signal elements by extracting said quadrature-related carriers, said received signal elements being different from said data signal elements due to the transformation of said carriers into a single-sideband signal; and

means for recovering said data signal elements by forming at least one estimate of each of said data signal elements by altering one of said received signal elements by at least one preselected quantity and then comparing said formed estimate against a preselected criterion.

2. The apparatus of claim 1 wherein said estimate formed for each of said data signal elements involves altering a different one of said received signal elements.

3. The apparatus of claim 2 wherein said recovery means forms said estimate at selected times.

4. The apparatus of claim 1 wherein each of said data signal elements have specific assigned values.

5. The apparatus of claim 4 where said assigned values for all data signal elements are the same.

6. The apparatus of claim 4 wherein said assigned values for one data signal element are different from said assigned values for any other data signal elements.

7. The apparatus of claim 4 wherein said preselected quantity is a function of the assigned values.

8. The apparatus of claim 7 wherein said function forms a set of numbers which comprises said preselected quantity.

9. The apparatus of claim 1 wherein each

received signal element includes one data signal element and a nonzero function of another data signal element.

10. The apparatus of claim 9 wherein said nonzero function is the Hilbert transform.

11. The apparatus of claim 4 wherein said preselected quantity lies in a set of numbers found by taking an algebraic combination of all possible permutations of said assigned values of one of said data signal elements.

12. Receiver apparatus for use in a digital transmission system wherein elements of a data signal modulate quadrature-related carriers and wherein said carriers are then transformed into a single-sideband signal, said receiver apparatus comprising

means for demodulating said single-sideband signal by extracting said quadrature-related carriers to form elements of a received signal, each received signal element including a selected element of said data signal and a nonzero function of an unselected data signal element; and

means for recovering said data signal elements by altering each received signal element by a plurality of preselected amounts so as to form a set of values for each received signal element and then picking one value from each set in accordance with a preselected criterion.

13. The apparatus of claim 12 wherein said nonzero function is the Hilbert transform of the unselected data signal element.

14. A method of retrieving elements of a data signal wherein said elements modulate quadrature-related carriers and wherein said carriers are transformed into a single-sideband signal, said method comprising the steps of

demodulating said single-sideband signal to form received signal elements by extracting said quadrature-related carriers, said received signal elements being different from said data signal elements due to the

transformation of said carriers into a single-sideband signal; and

recovering said data signal elements by forming at least one estimate of each of said data signal elements by altering one of said received signal elements by at least one preselected quantity and then comparing said formed estimate against a preselected criterion.

15. A transmitter for use in communication systems comprising

means for modulating quadrature-related carrier signals with elements of a data signal to form a double-sideband signal, and

means for transforming said double-sideband signal into a single-sideband signal.

16. A communications system comprising a transmitter and a receiver wherein said transmitter comprises

means for modulating quadrature-related carrier signals with elements of a data signal to form a double-sideband signal, and

means for transforming said double-sideband signal into a single-sideband signal, and

said receiver comprising

means for demodulating said single-sideband signal to form received signal elements by extracting said quadrature-related carriers, said received signal elements being different from said data signal elements due to the transformation of said carriers into a single-sideband signal; and

means for recovering said data signal elements by forming at least one estimate of each of said data signal elements by altering one of said received signal elements by at least one preselected quantity and then comparing said formed estimate against a preselected criterion.

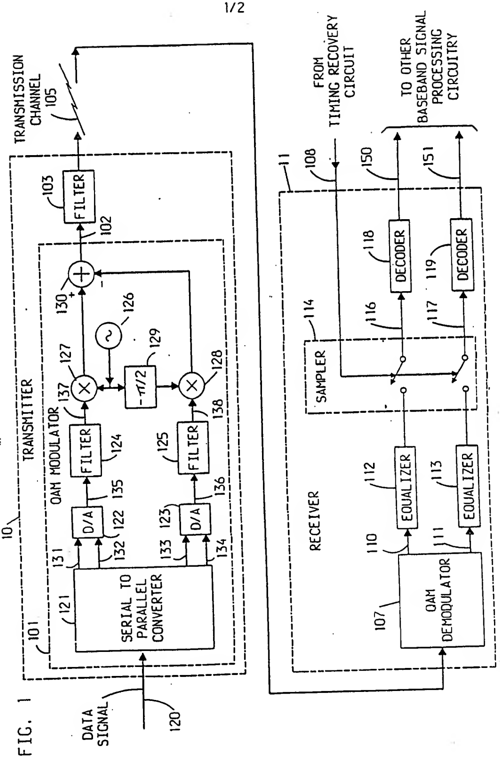

TRANSMISSION CHANNEL
105
TRANSMITTER
10
QAM MODULATOR
101
SERIAL TO PARALLEL CONVERTER
121
131
132
133
134
D/A
122
123
135
136
FILTER
124
125
137
138
127
128
126
$-\pi/2$
129
130
+
-
102
FILTER
103
DATA SIGNAL
120
RECEIVER
11
QAM DEMODULATOR
107
110
111
EQUALIZER
112
113
SAMPLER
114
116
117
DECODER
118
119
150
151
108
FROM TIMING RECOVERY CIRCUIT
TO OTHER BASEBAND SIGNAL PROCESSING CIRCUITRY

FIG. 1

FIG. 2

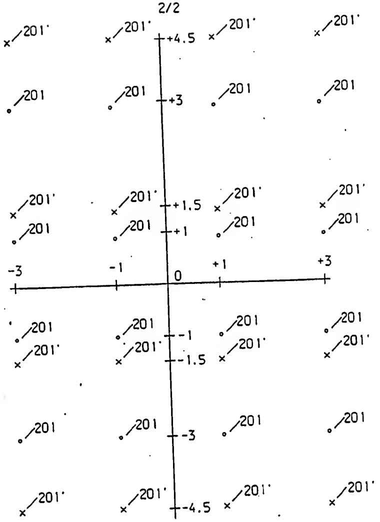
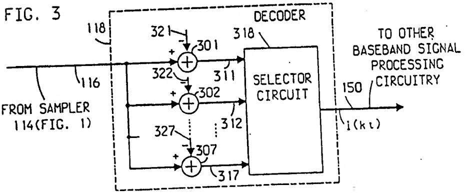

201'
201
+4.5
+3
+1.5
+1
-3
-1
0
+1
+3
-1
-1.5
-3
-4.5

FIG. 3


DECODER
118
321
301
318
322
302
311
SELECTOR
CIRCUIT
116
FROM SAMPLER
114(FIG. 1)
327
312
307
317
TO OTHER
BASEBAND SIGNAL
PROCESSING
CIRCUITRY
150
i(kτ)

INTERNATIONAL SEARCH REPORT

International Application No PCT/US85/00302

I. CLASSIFICATION OF SUBJECT MATTER (if several classification symbols apply, indicate all) [3]

According to International Patent Classification (IPC) or to both National Classification and IPC

Int. CL.[3] H04L 25/03 27/02
U.S. CL. 375/39, 43, 77

II. FIELDS SEARCHED

Minimum Documentation Searched [4]

Classification System	Classification Symbols
U.S.	375/39,43,77,80,81 370/20 329/112

Documentation Searched other than Minimum Documentation to the Extent that such Documents are Included in the Fields Searched [5]

III. DOCUMENTS CONSIDERED TO BE RELEVANT [14]

Category *	Citation of Document, [16] with indication, where appropriate, of the relevant passages [17]	Relevant to Claim No. [18]
Y,P	US, A, 4,470,145 04 September 1984 Williams	1-16
X,P	US, A, 4,461,011 17 July 1984 Lender et. al.	15
Y	US, A, 4,439,863 27 March 1984 Bellamy	1-14, 16
X	US, A, 3,605,017 14 September 1971 Chertok et. al.	15
Y	US, A, 3,522,537 04 August 1970 Boughtwood	1-16
X	US, A, 3,443,229 06 May 1969 Becker	15
A	US, A, 3,849,730 19 November 1974 Ho	1-16

* Special categories of cited documents: [15]
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"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.
"&" document member of the same patent family

IV. CERTIFICATION

Date of the Actual Completion of the International Search [2]	Date of Mailing of this International Search Report [2]
12 March 1985	27 MAR 1985
International Searching Authority [1]	**Signature of Authorized Officer** [10]
ISA/US	Stephen Chin

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(54) Block coded moduletion system.

(57) A femily of block coded modulation systems for achieving coding geins using eesily implemented coding and decoding methods is described.

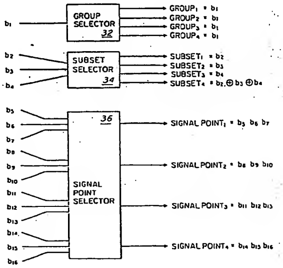

FIG. 4

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BLOCK CODED MODULATION SYSTEM

Background of the Invention

This invention relates to high-speed data transmission over band-limited channels, such as telephone lines.

Modulation systems for such channels commonly use two-dimensional carrier modulation, generically called double-sideband-quadrature-carrier (DSB-QC) modulation. Such modulation systems are discussed, for example, in U.S. Patent 3,887,768 (Forney/Gallager), incorporated herein by reference, which also shows an implementation of such a system.

Conventional DSB-QC systems are used to send an integer number (N) of bits in each modulation interval of T seconds in a nominal bandwidth of 1/T Hz. For example, some telephone line modems send 4 bits per modulation interval of 1/2400 sec. within a nominal bandwidth of 2400 Hz, thus achieving a 9600 bits per second (bps) data transmission rate. Modems for sending 6 bits per modulation interval to achieve 14,400 bps in a nominal 2400 Hz bandwidth are also available.

Such DSB-QC systems use signal constellations of 2^N signals for sending N bits per modulation interval. A family of such constellations with signals arranged on a regular rectangular grid is described in Campopiano and Glazer, "A Coherent Digital Amplitude and Phase Modulation Scheme," IRE Transactions on Communication Systems, Vol. CS-9, pp. 90-95, March 1962, incorporated herein by reference.

Fig. 1 shows the arrangement of signals in the Campopiano and Glazer constellations and the outer boundaries of those constellations for values of N from 4 through 8. Table I shows the so-called "required signal-to-noise ratio" $\overline{P}$ (defined in Campopiano et al. and in the Forney/Gallager patent) for the

constellations of Fig. 1. Each unit increase in N corresponds to an increase in required signal-to-noise ratio of about a factor of two or 3.0 decibels (dB).

Table I (Campopiano and Glazer)

N	S	$\overline{P}$	(dB)
4	16	10	10.0
5	32	20	13.0
6	64	42	16.2
7	128	82	19.1
8	256	170	22.3

N = number of bits sent per modulation interval

S = number of signals in signal constellation

$\overline{P}$ = required signal-to-noise ratio

(dB) = $\overline{P}$ measured in decibels

For higher transmission speeds, so-called coded modulation techniques can provide improved resistance to noise and other channel impairments; that is, they can provide a reduction in required signal-to-noise ratio, or a so-called "coding gain".

In such coded systems, to send N bits per modulation interval, signal constellations having more than 2^N signals are used, and coding is used to introduce dependencies between modulation intervals so that the set of available signals from which a signal point can be selected in one modulation interval depends in general on the signal points selected for other modulation intervals.

In one coding technique for getting a coding gain (disclosed in Csajka et al, U.S. Patent 4077021, and Ungerboeck, "Channel Coding with Multilevel/Phase Signals," IEEE Transactions on Information Theory, Vol. IT-28, pp. 55-67, January, 1982), the N bits appearing in each modulation interval are individually mapped into signal points selected from a constellation of 2^{N+1}

signals. The signals in the constellation are organized into subsets such that the minimum distance between two signals belonging to one subset is greater than the minimum distance between any two signals in the constellation. The selection of the signal point for each N input bits is made to depend, in part, on the historical sequence of all previously selected signal points, as represented by the state of a finite state device in the encoder. This so-called trellis coding effectively permits only certain sequences of signal points to be transmitted, and the coded historical information carried by every signal point is exploited at the receiver by a maximum likelihood sequence estimation technique (e.g., one based on the Viterbi algorithm, as described in Forney, "The Viterbi Algorithm," Proceedings of the IEEE, Vol. 61, pp. 268-278, March, 1973, incorporated herein by reference).

A coding gain can also be realized using a block coded modulation system in which blocks of n input bits are sent in m modulation intervals, so that $N=n/m$ bits are sent per modulation interval. For each block, m signal points are selected from a constellation having more than 2^N signals, a process which is equivalent to mapping each block into a code word selected from an available code word set of 2^n code words arranged in 2m-dimensional space (called simply 2m-space), with the 2m coordinates of each code word, taken two at a time, defining the respective pair of coordinates in two-dimensional space of the m signal points to be selected. The code word set from which the code word for any block may be drawn is independent of the signal points selected for any other block. At the receiver, the decisions on which signal points were sent are based

on the received signal points for each block (the so-called received word), preferably using maximum likelihood decoding.

One method of block coding involves using a code word set arranged on a finite portion of a densely packed infinite geometrical lattice in 2m-space; see, for example, Conway and Sloane, "Fast Quantizing and Decoding Algorithms for Lattice Quantizers and Codes," IEEE Transactions on Information Theory, Vol. IT-28, pp. 227-232, March 1982, and the references cited therein, incorporated herein by reference. A representative system of this type is disclosed in "Block Coding for Improved Modem Performance," a Canadian contribution (Com XVII-No. 112) to Study Group XVII of the International Telegraph and Telephone Consultative Committee (C.C.I.T.T.), March 1983, incorporated herein by reference, which describes an 8-space code for sending 4 bits per modulation interval with an asymptotic coding gain of 3.4 dB over the uncoded Campopiano and Glazer 16-signal constellation (defined by the N=4 boundary in Fig. 1).

Summary of the Invention

The invention includes a family of block coded modulation systems of progressively greater complexity exhibiting progressively greater coding gain over uncoded systems.

In general, in one aspect the invention features, in such block coded modulation systems, the improvement in which the signal constellation includes groups having equal numbers of signals, and the encoder is arranged so that the group from which at least one signal point for a block is drawn depends on the group from which at least one other signal point for the block is drawn, whereby a coding gain over uncoded systems is achieved.

In preferred embodiments, the signals are arranged (e.g., on a rectangular grid) so that the minimum squared distance between two signals belonging to the same group is greater than (e.g., twice the distance) between two signals belonging to different groups; the signal points are two dimensional, the code words are 2m-dimensional, m being the number of signal points corresponding to each code word, and each block of digital data has mN bits; each block comprises a plurality of bits, and the groups from which the signal points for the block are drawn are determined by a single bit of the block; N is an integer, m is an even integer no smaller than 2, and the constellation has $1.5x2^N$ signals; N=r+1/2, r being an integer, m is no smaller than 2, and the constellation comprises 2^{r+1} signals; the encoder is further arranged so that m-1 signal points for each block may be drawn from among all signals in the constellation, and the one remaining signal point for the block is drawn from a group that depends (e.g., by means of a single-parity-check code or a state-transition trellis) on the groups from which the m-1 signal points are drawn; m is 2; and N is 4 or 4 1/2.

In other embodiments, the groups each have subsets having equal numbers of signals, and the encoder is further arranged so that the subset from which at least one signal point for the block is drawn depends on the subset from which at least one other signal point for the block is drawn; the signals are arranged (e.g., on a rectangular grid) so that the minimum squared distance between two signals belonging to one subset is greater than (e.g., twice the distance) between two signals belonging to different subsets within the same group; the subsets from which the signal points for a block are drawn are determined based on a plurality of

bits representing less than all of the digital data of the block; N is an integer, m is at least 4, and the constellation has 2^{N+1} said signals; N=r+1/2, r being an integer, m is an even integer no smaller than 4, and the constellation has $1.5x2^{r+1}$ signals; the encoder is further arranged (e.g., by means of a single-parity-check code, a Hamming code, or a state-transition trellis) so that one signal point for each block may be drawn from among all signals in the constellation, m-2 of the signal points are drawn from groups that depend on the group from which that one signal point is drawn, and the single remaining signal point for the block is drawn from a subset that depends on the subsets from which the one signal point and the m-2 signal points are drawn; m is 4; and N is 4.

In other embodiments, the subsets each comprise classes having equal numbers of signals, and the encoder is further arranged so that the class from which at least one signal point for the block is drawn depends on the class from which at least one other signal point for the block is drawn; the minimum squared distance between two signals belonging to one class is greater than between two signals belonging to different classes within the same subset; each block has a plurality of bits, and the classes from which the signal points for the block are drawn are determined based on a plurality of bits representing less than all of the digital data of the block; m is 8; N is an integer and the constellation has $1.5x2^{n+1}$ signals; N=r+1/2, r being an integer, m is 8, and the constellation comprises 2^{r+2} signals.

In other embodiments, the classes each comprise subclasses having equal numbers of signals, and the encoder is further arranged so that the subclass from

which at least one signal point for the block is drawn depends on the subclass from which at least one other signal point for the block is drawn; the minimum squared distance between two signals belonging to one subclass is greater than (e.g., twice the distance) between two signals belonging to different subclasses within the same class; m is 12; N is an integer and the constellation has 2^{N+2} signals; the subclasses from which the signal points for the block are drawn are determined by the encoder based on a Golay code applied to at least one of the bits; m is 12 and N is 4; and the subclasses from which the signal points for the block are drawn are determined based on a plurality of bits representing less than all of the digital data of the block.

In other embodiments, there are means for selecting at least one signal point of a block on the basis of less than all digital data in the block; the constellation has quadruplets each having four signals located at the same distance from the origin but separated by 90° intervals about the origin, the digital data comprises bits, and at least two of the bits are quadrantally differentially encoded; the four signals belonging to each quadruplet are drawn from four different subsets; there is also a decoder arranged to decide which code word was sent based on maximum likelihood sequence estimation in accordance with the Viterbi algorithm; there is a demodulator having a decoder arranged to make tentative decisions about which signal point was sent prior to final decoding of all of the received signal points for a block, and the demodulator comprises adaptive control circuitry arranged to be responsive to the tentative decision; the signal points for each code word are drawn from the same

group; the decoder is arranged to decide which code word was sent by first making a separate tentative decision based on code words whose signal points are drawn from each group, and thereafter a final decision based on the separate tentative decisions; selection of signal points for a block depends on a single-parity-check code based on at least one of the bits of the digital data; and the decoder is arranged to decide which code word was sent by first making a tentative decision as to each signal point in the code word without regard to whether the parity check is satisfied, and accepting the tentative decisions as the final decision, either without changes if the parity check is satisfied, or after changing the least reliable one of the tentative decisions if the parity check is not satisfied.

In another aspect, the invention features, in a modulation system for sending a block of digital data bits over a band-limited channel using a plurality of modulation signal points drawn from a two-dimensional constellation of available signals, the improvement in which the constellation has a plurality of inner signals, and a plurality of outer signals further from the origin than the inner signals, one bit of the digital data determines whether any of the plurality of signal points will be drawn from the outer signals, and if an outer signal will be drawn, at least one other bit of the digital data determines which of the plurality of signal points will be an outer signal point.

In preferred embodiments, there are 2^t signal points, the digital data comprises at least t+1 bits, there are S inner signals and $2^{-t}S$ outer signals, and t bits determine which signal point will be an outer signal point; t=1; the block has 2N+1 bits and S is 2^N, N being an integer; N-1 of the bits determine

which outer signal is drawn, and N bits determine which inner signal is drawn; N is 4; t is 2; each block has 4N+1 bits and S is 2^N, N being an integer; and N-2 of the bits determine which outer signal is drawn and N bits determine which inner signal is drawn.

The 4-space, 8-space, 16-space, and 24-space modulation systems of the invention respectively exhibit asymptotic coding gains of about 1.5, 3.0, 4.5, and 6.0 dB over uncoded systems that send the same number of bits per modulation interval through the same nominal bandwidth using DSB-QC modulation. The block coded systems of the invention are slightly inferior in performance to the best known prior art schemes (e.g., about .2 dB and .4 dB worse for 4-space and 8-space systems, respectively), but have substantial advantages in implementation. The invention uses constellations having relatively small numbers of signals. The encoding techniques are economical, simple, and easily implemented. Encoding delay is reduced because early signal points selected for a given block can be sent before encoding of the block has been completed. Decoding delay is small because received signal points can be finally decoded as soon as all signal points for a given block have been received. Useful tentative decoding decisions can be made even sooner. Non-integral numbers of bits per interval can be sent. Quadrantal differential encoding can be used to improve immunity to phase hits. Economical, simple, and easily implemented decoding techniques are made possible.

Other advantages and features of the invention will be apparent from the following description of the preferred embodiment, and from the claims.

Description of the Preferred Embodiment

We first briefly describe the drawings.

Drawings

Fig. 1 is a diagram of a family of five prior art signal constellations.

Fig. 2 is a block diagram of modem apparatus.

Fig. 3 is a diagram of a signal constellation in accordance with the preferred embodiment.

Fig. 4 is a block diagram of the signal selection logic for the constellation of Fig. 3.

Fig. 5 is a table showing the correspondence between input bits and signal subsets.

Fig. 6 is a state-transition diagram (trellis) for use with the constellation of Fig. 3.

Fig. 7 is a diagram of a family of four signal constellations in accordance with an alternate embodiment.

Figs. 8, 12, 15, 17, and 20 are block diagrams of signal selection logic in accordance with alternate embodiments.

Figs. 9, 10, 11, 14, 16, and 19 are diagrams of signal constellations in accordance with alternate embodiments.

Figs. 18 and 21 are diagrams of the relationships of signal quadruplets in the constellations of Figs. 16 and 19, respectively.

Figs. 13, 22, and 23 are trellises for use with alternate embodiments.

In the following section, the preferred embodiment (an 8-space system sending 4 bits per modulation interval and achieving an asymptotic coding gain of 3.0 dB) is described first. Alternative embodiments thereafter described are other decoding and encoding techniques for 8-space block coded modulation

systems; 8-space block coded systems for sending any integral number of bits per modulation interval; multidimensional signal structures for sending a non-integral number of bits per modulation interval (useful in both coded and uncoded systems); 4-space, 16-space, 24-space, and 2m-space block coded systems; and other embodiments.

Structure and Operation

In the preferred embodiment (an 8-space block coded modulation system for sending 4 bits per modulation interval), a block of 16 bits is sent in each 4 modulation intervals using a signal constellation having 32 signals.

Referring to Fig. 2, in transmitter 10 the bits appearing in an input bit stream 12 are grouped by serial/parallel converter 14 into 16-bit blocks 16 (i.e., n=16). Each block 16 is encoded by signal point selection logic 18 into a sequence of four (i.e., m=4) 2-dimensional signal points 20 (which are then used for conventional DSB-QC modulation by modulator 22 for transmission over channel 24).

Referring to Fig. 3, the signal points are selected from a two-dimensional signal constellation 30 having 32 signals (i.e., twice the number of signals (16) needed for uncoded modulation at 4 bits per modulation interval, as with the signal constellation shown within the N=4 boundary of Fig. 1). The 32 signals are arranged in a rectangular grid with integer coordinates, each signal point having one odd and one even coordinate (in an arrangement like the Campopiano and Glazer constellation shown within the N=5 boundary of Fig. 1, but rotated by 45°) and are divided into four disjoint subsets (A_0, A_1, B_0, B_1) each having eight signals. The four subsets A_0, A_1, B_0, B_1

are arranged in two groups (A, comprising subsets A_0 and A_1, and B, comprising B_0 and B_1) of two subsets each. The arrangement of signals in the subsets and groups is such that the minimum squared distance ($d_o^2=2$) between any two signals on the plane (e.g., between an A signal and a B signal) is smaller than the minimum squared distance ($2d_o^2=4$) between any two signals in one group (e.g., between an A_0 signal and an A_1 signal), which is in turn smaller than the minimum squared distance ($4d_o^2=8$) between any two signals in the same subset (e.g., between an A_0 signal and another A_0 signal).

Signal selection logic 18 is arranged to select, for each block of input bits, a sequence of four signal points which are interdependent, i.e., the signal group and/or subset from which at least some of the signal points for a given block may be selected depends in part on at least one of the other signal points selected for that block. The code word set from among which the code word for a given block may be drawn is independent of the signal points selected for any other block.

Each of the 32 constellation signals can be uniquely specified by a total of 5 bits: one bit naming the group from which it is taken, one bit naming the subset within that group, and three bits naming which of the eight signal points within the named subset is to be selected. Thus the four signal points for a block could be uniquely specified by a total of 20 bits. However, the block has only 16 input bits on which to base the selections.

Referring to Fig. 4, the group, subset, and signal point selection information needed to select each of the four signal points for each block is derived from the 16 input bits (b_1-b_{16}) as follows.

Group selector 32 uses bit b_1 (called the group bit) to determine the one group from which all four signal points for a given block is to be drawn. (The designation $group_1$ refers to the group from which the first signal point is selected; and so on.) $Group_1$ through $group_4$ are therefore always the same group for a given block. For example, if $b_1=0$ all four signal points would be selected from group A; if $b_1=1$, all would be from group B.

Subset selector 34 uses bits b_2, b_3, and b_4 to determine from which subsets of the selected group each of the four signal points is to be drawn. Bits b_2, b_3, and b_4 respectively determine $subset_1$, $subset_2$, and $subset_3$. For example, in a block for which the signal points are from group A, if $b_2=0$ the first signal point would be selected from subset A_0, otherwise (if $b_2=1$) from A_1. $Subset_4$ is a parity bit generated by subset selector 34 such that bits b_1, b_2, b_3 and the parity bit (called subset bits) include an even number of 1 bits. Thus the subset bits can be viewed as coded bits derived by a (4, 3) single-parity-check code from the input bits b_2-b_4. The remaining bits (b_5-b_{16}) taken three at a time (called signal point bits) specify which particular four signal points ($signal\ point_1$-$signal\ point_4$) are to be drawn by signal point selector 36 from the sele ted 8-point subsets. Thus group selector 32 and subset selector 34 assure that the group from which at least one of the signal points for a given block is selected depends on the group from which at least one of the other signal points for the block is selected (thereby reducing, after the first signal point, the number of groups from which later signal points for a block can be selected), and to assure that

the subset from which at least one of the signal points is selected depends on the subset from which at least one of the other signal points is selected (thereby reducing, after the first signal point, the number of subsets from which later signal points for a block can be selected).

Although the signal selection (coding) has been explained as the selection of four successive signal points from a 2-dimensional signal constellation, it could also be viewed as an 8-space block coded modulation technique in which a set of 2^{16} code words are arranged on a lattice in 8-space, and coding consists of selecting one of the code words for each block. The eight coordinates of the selected code words could then be taken two at a time to specify the four two-dimensional signal points to be used for carrier modulation.

The required signal-to-noise power ratio for the 32-point signal constellation of Fig. 3 is $\overline{P}=10$, the same as for the N=4 signal constellation of Fig. 1. The minimum squared distance between code words in 8-space is 8, double the minimum squared distance (4) between signals in the N=4 Fig. 1 constellation, providing an asymptotic coding gain of a factor of 2, or 3.0 db.

That the minimum squared distance between code words in 8-space is 8 can be seen as follows. Two different code words which have the same group bit and the same subset bits must differ in at least one of their respective four signal points. Because those two different signal points must be drawn from the same subset, and signal points in the same subset have a minimum squared distance of 8 on the two-dimensional plane, the two code words in 8-space likewise have a minimum squared distance of 8.

Similarly, two code words which have the same group bit but different subset bits must differ in at least two of their subset bits (because the subset bits have even parity). At least two of their respective four signal points must therefore differ in each case by a minimum squared distance of 4 (the minimum squared distance on the two-dimensional plane between two signal points chosen from a given group) for a total minimum squared distance of 8.

Likewise, two code words which have different group bits differ in all four of their signal points, and the minimum squared distance between two signal points on the two-dimensional plane is 2 for a total minimum squared distance of 8. Thus, in all possible cases, the minimum squared distance is 8.

The theoretical coding gain of 3.0 dB is partially offset by the effect of the error event probability coefficient (see Forney Viterbi article), which can be estimated as follows. In 8-space, each code word has up to 240 nearest neighbors (16 which have the same group and subset bits, 96 which have the same group bits but different subset bits, and 128 which have different group and subset bits). But because code words near the outer boundary of the code word set have fewer neighbors, the average code word only has about 180 near neighbors. The error event probability coefficient per modulation interval is therefore of the order of 180/4 = 45, corresponding to a fraction of a decibel of loss in performance for error probabilities of interest.

At the receiver, decoding of the four signal points into the corresponding bit stream is optimally done by maximum likelihood decoding, i.e., picking the

one code word which is most likely to have been sent, given the 8-space received word determined by the four received two-dimensional signal points.

For each of the received two-dimensional signal points, the decoder tentatively finds (by conventional slicing in two-dimensional space) the closest (in Euclidean distance) signal from the A group and the closest signal from the B group to the received signal point, noting, on a running basis, which of the coordinates of the tentatively selected signals for each group is least reliable (i.e., has the greatest apparent error). The subset bits of the four tentatively selected A signals are checked for parity. If parity checks, then the four tentatively selected A signals define the closest code word in the A group to the received word. If parity fails to check, the closest A code word is found by changing the least reliable coordinate to the coordinate of the next closer A signal, which (because of the arrangement of the subset points in the constellation) will always cause a change of that signal from an A_0 to an A_1 signal (or vice versa), and thus yield a code word of correct parity. An analogous slicing and parity checking sequence finds the closest B code word to the received word.

The final decoding decision entails choosing whether the previously determined closest A code word or closest B code word is closer (in overall Euclidean distance) to the received coordinates in 8-space. The choice may be made by computing the sum of the squared coordinate differences between each of the two code words and the received word and comparing the sums.

Decoding thus requires only slicing (twice) in each coordinate, storing (for A and B code words respectively) the location and magnitude of the largest

apparent error so far, accumulating the Euclidean distances for the best code words from each group, checking the parity of each best code word (and changing one coordinate in each code word, if needed, thereby changing the Euclidean distance for that word), and comparing the Euclidean distances for the resulting two best code words to select the better one. This decoding method may be readily and simply implemented (by conventional techniques) using a programmable microprocessor.

The block coded modulation system can be made transparent to 90° phase rotations between the received carrier and the transmitted carrier using quadrantal differential coding on a block basis. The signal constellation of Fig. 3 is arranged in quadruplets of signals having the same radii but separated by 90° phase differences, each quadruplet having one signal from each subset, so that a 90° clockwise rotation translates, within each quadruplet, the A_0 signal to the B_0 signal, the B_0 signal to the A_1 signal, the A_1 signal to the B_1 signal, and the B_1 signal to the A_0 signal; and analogous translations occur for 180° and 270° rotations. For example, the signals denoted 37 in Fig. 3 form such a quadruplet. Referring to Fig. 5, if each signal's subset and group bits are taken together as a 2-bit integer, 90° clockwise rotation of any signal corresponds to an increment of 1 (modulo 4) in the value of that integer. Quadrantal differential coding can then be accomplished by assigning the same three signal point bits to all four of the signals within each quadruplet, and coding the subset and group bits for each signal point in a code word by differential four-phase coding, using for example the subset and group bits of the last signal point of the previous code word as the reference.

If (x_1, x_2) denotes the differentially encoded subset and group bits of the last signal point of the previous code word, and (y_1, y_2) denotes the subset and group bits of a signal point in the current code word (before differential encoding), then (z_1, z_2), the differentially encoded subset and group bits of that signal point in the current code word, is determined by adding (x_1, x_2) to (y_1, y_2) as two-bit integers, modulo four. Because $z_2=x_2 \oplus y_2$, the same differentially encoded group bit is used for all signal points in the block. At the receiver, assume that both the previous and the current code words are correctly decoded (except for a possible phase rotation of an integral multiple of 90°). If the received bits are denoted (x_1', x_2') and (z_1', z_2'), then (y_1, y_2) is determined by subtracting (x_1', x_2') from (z_1', z_2') as two-bit integers, modulo four. The resulting (y_1, y_2) value will then be correct even if both (x_1', x_2') and (z_1', z_2') have been advanced by p modulo 4 as a result of a p x 90° phase rotation.

Because decoding of a code word can be finished as soon as all four received signal points are determined, the decoding delay and error propagation of this block coding system are strictly limited to four modulation intervals. Further, because tentative signal point decisions (made before all four signal points have been received) are only about 3 dB less reliable than in the uncoded 16-signal constellation of Fig. 1, those tentative decisions may be used to update adaptive equalizers and other tracking loops in the receiver without even waiting for all four signal points to be received. The tentative decisions are made by finding the closest signal on the constellation of Fig. 3 to

each received signal point, which can be accomplished by rotating the coordinates 45° with respect to the signal constellation, followed by conventional slicing in each of the two rotated coordinates. At the transmitter, transmission can begin as soon as five bits (b_1, b_2, b_5, b_6, b_7) have been collected, thereby reducing the encoding delay which would exist for other kinds of block coding in which all of the bits of the block would have to be collected before encoding could begin.

Other Embodiments

Other Decoding and Encoding Techniques for 8-Space Block Coded Systems.

In another embodiment, maximum likelihood decoding can be accomplished by applying the Viterbi algorithm (in a manner described in Forney's Viterbi article, cited above).

Referring to Fig. 6, in order to apply the Viterbi algorithm, the encoding process can be illustrated by a finite-length 4-state trellis 40. The trellis shows from left to right the state of the encoder at five successive times, beginning with a node 42 reflecting the time immediately before encoding of a new block begins and ending with a node 44 reflecting the time just after the encoding of that block has been completed. In between nodes 42, 44 are shown four modulation intervals separated by three time instants, each such instant being represented by a column of four nodes 46 corresponding to the four states which the encoder can occupy at that instant. Branches 48 from each node to the next nodes to its right correspond to modulation intervals and reflect transitions of the state of the encoder which can occur as a signal point is selected. Each branch 48 is labeled with a letter to

indicate the subset corresponding to the signal point associated with that modulation interval. Each node 46 is marked with a two-bit value, the first bit reflecting the group bit for all signal points for that block (i.e., 0 = A, 1 = B), and the second bit reflecting the value of the accumulated parity check subset bit as of that time.

The encoder begins at a common node 42 (reflecting that the encoder's state does not depend on historical information). From node 42 the encoder branches to one of four states depending on the group and subset bits of the first selected signal point. For example, if the first selected signal point is from subset A_0, the next encoder state is at node 50. For the next two signal points, the encoder state stays in the half of the trellis defined by the group bit of the first signal point, but switches states (from one parity to the other) in accordance with the transitions shown. Only 16 of the 32 signals in the constellation are available for selection of the second and third signal. The transition to the final node 44 is determined by the parity state of the encoder (e.g., an A_1 or B_1 point is sent if the parity state is 1, otherwise an A_0 or B_0 point is sent). Only a subset of 8 of the 32 signals in the constellation are available for selection of the fourth signal point, and that subset depends on the first three selected signal points. Thus the paths through the trellis define the possible combinations of subsets from which the sequence of four signal points for each block can be chosen.

In decoding each received word, in accordance with the Viterbi algorithm, decisions are first made (using conventional slicing in two dimensions, with the coordinate axes rotated 45°) of which of the eight

signals in each of the four subsets is closest to each received signal point. A branch in trellis 40 (corresponding to each subset) is then associated with the closest signal in its corresponding subset, and the so-called metric of each branch is the squared distance between the received signal point and the corresponding closest signal associated with that branch. The remaining steps of the Viterbi algorithm decoding proceed as described in Forney's Viterbi article, cited above. A final decoding decision is possible after four modulation intervals, at which time all trellis branches converge to a single node 44.

In another embodiment, the input bits may be coded (using a Hamming code) into coded bits which are used to select the subsets from which the signal points may be selected.

Four bits in each block, b_1-b_4, rather than being used in a group selector and subset selector (as shown in Fig. 4) to select the group and subset from which each signal point is to be selected, are encoded into 8 coded bits z_1-z_8 using the following conventional extended (8, 4) binary Hamming code (Hamming codes are discussed, e.g., in Berlekamp, Algebraic Coding Theory, McGraw-Hill, 1968, incorporated herein by reference):

$$
\begin{aligned}
z_1 &= b_2 \\
z_2 &= b_1 \oplus b_2 \\
z_3 &= b_3 \\
z_4 &= b_1 \oplus b_3 \\
z_5 &= b_4 \\
z_6 &= b_1 \oplus b_4 \\
z_7 &= b_2 \oplus b_3 \oplus b_4 \\
z_8 &= b_1 \oplus b_2 \oplus b_3 \oplus b_4
\end{aligned}
$$

which can be generated using, for example, the following generator matrix (having minimum Hamming distance 4):

01010101
11000011
00110011
00001111

Bits $z_1 - z_8$ are taken in pairs to select the subsets of Fig. 3 from which the four signal points are to be drawn in accordance with the following table:

Bit Pair	Subset
00	A_0
01	B_0
10	B_1
11	A_1

The Hamming distance (d_H) between bit pairs is then equal to half the minimum squared distance between signals in the corresponding subsets, i.e., $d^2(A_0, B_0) = d^2(A_0, B_1) = d^2(A_1, B_0) = d^2(A_1, B_1) = 2$, corresponding to $d_H(00, 01) = d_H(00, 10) = d_H(11, 01) = d_H(11, 10) = 1$; and $d^2(A_0, A_1) = d^2(B_0, B_1) = 4$, corresponding to $d_H(00, 11) = d_H(01, 10) = 2$.

The minimum squared distance between code words in 8-space is then 8. For if two code words correspond to different Hamming coded bits z_1-z_8, then these bits differ in at least 4 places, which (because the Hamming distance is half the subset distance, as just described) implies that the squared distance between the code words is at least 8. Alternatively, two code words corresponding to the same Hamming coded bits z_1-z_8 have signal points all in the same subsets in each modulation interval; in at least one interval the signal points must be different, and since they belong to the same subset the minimum squared distance between them is 8.

In other embodiments, the selection of the four subsets to be used in the four modulation intervals can be done by any method (in addition to the three already described) which produces minimum squared distance of 8 between code words (based on the distance properties of the subsets in 2-space). The 12 bits used to select the particular signal point for each modulation interval (or the original 16-bit block) can be arbitrarily transformed. The constellation can be any signal constellation that can be divided into four eight-point subsets with the required distance properties.

<u>8-Space Block Coded Systems for Sending Any Integral Number of Bits per Modulation Interval</u>

In other embodiments, any number of bits (N) per modulation interval can be sent by using a signal constellation on a rectangular grid having 2^{N+1} signals (e.g., constellations such as the ones shown in Fig. 1), divided into four equal subsets (each having 2^{N-1} signals) and arranging the signals so that alternate signals fall respectively in groups A and B, and alternate signals within each group fall respectively in subsets (A_0, A_1; B_0, B_1). Four bits of each block of 4N bits are used to select the four subsets by any method (including those described above) which assures a minimum squared distance between code words of $4d_o^2$ (where d_o^2 is the minimum squared distance between signals in the constellation). The remaining 4 (N-1) bits are used to select which signals within the selected subsets are to be sent in each interval.

Table II shows the required signal-to-noise ratios $\overline{P}$ for values of N from 3 through 7 using the constellations of Fig. 1 with S signal points and the 8-space block coding system described above:

Table II

N	S	$\bar{P}$	(dB)
3	16	2.5	4.0
4	32	5	7.0
5	64	10.5	10.2
6	128	20.5	13.1
7	256	42.5	16.3

Compared with the uncoded systems treated in Table I, about 3 dB of coding gain is achievable for any integer value of N, using a signal constellation of size 2^{N+1} rather than 2^N.

Multidimensional Signal Structures for Sending a Non-Integral Number of Bits per Modulation Interval

In other embodiments, signal systems of multiple (at least four) dimensions are used for sending a non-integral number (N) of bits per modulation interval, where N is of the form N=r+f (r an integer, f a fraction between 0 and 1). Such systems send blocks of bits (each block longer than one modulation interval), and are efficient for use in uncoded systems (as well as coded systems) because their required signal-to-noise ratio is only about fx3 dB more than for a comparable Fig. 1 signal constellation for sending r bits per modulation interval, and because the number of signals in the constellation is about $(1+f)2^r$.

Referring to Fig. 7, signal constellations for use in sending N=r+1/2 bits per modulation interval each comprise the 2^r signals in the N=r constellation of Fig. 1 (called inner signals) plus an additional 2^{r-1} signals (called outer signals) generally arranged further from the origin, on the same grid, and with the same symmetries, as the inner signals.

To send blocks of 2r+1 bits in two modulation intervals, a first bit of the 2r+1 bits determines whether any outer signal is to be sent. If not, the remaining 2r bits taken r at a time determine the two inner signals for the two intervals. If so, a second bit determines whether the outer signal will be sent in the first or second modulation interval, r-1 bits determine which outer signal, and the remaining r bits determine which inner signal is sent in the other interval. The average power required to send both signal points will be 3/4 the average power required to send inner signals (which are sent on average 3/4 of the time) plus 1/4 the average power required to send outer signals (which are sent on average 1/4 of the time).

For example, 4 1/2 bits per modulation interval (r=4) can be sent in 2 modulation intervals using the constellation having 24 signals (within the N=4 1/2 boundary of Fig. 7) divided into an inner group of 16 signals, arranged as shown in Fig. 1 (N=4) and an outer group of 8 signals, with the arrangement of signals exhibiting quadrantal symmetry.

Referring to Fig. 8, in the signal selection logic, bit b_1 is used by outer point selector 60 to determine whether none or one of the two selected signal points will be an outer signal. If one will be, then bit b_2 is used by selector 60 to determine which interval it will be in, and bits b_3-b_9 are used by signal point selector 62 to select the outer and inner signals as shown in Fig. 8. If neither signal point is to be an outer signal, then bits b_2-b_5 and b_6-b_9 are used by signal point selector 62 to select respectively the two inner signals.

The average power required to send inner points is 10.0, and to send outer points 26.0. The required signal-to-noise ratio is therefore 14 (11.5 dB), which is 1.5 dB more than the requirement for sending 4 bits per interval using the 16-signal constellation of Fig. 1.

Table III shows the numbers of constellation signals S and the required signal-to-noise ratios $\overline{P}$ for sending half-integer numbers N (between 4 1/2 and 7 1/2) of bits per interval using this system with the constellations of Fig. 7:

Table III

N	S	$\overline{P}$	(dB)
4 1/2	24	14	11.5
5 1/2	48	28.5	14.5
6 1/2	96	57	17.6
7 1/2	192	116	20.6

In other embodiments, any number of bits per interval of the form $N=r+2^{-t}$ (t an integer) can be sent by adding to inner signals (consisting of the 2^r signals in an uncoded signal constellation), outer signals (consisting of 2^{r-t} signals located as close to the origin as possible consistent with maintaining an overall rectangular grid exhibiting quadrantal symmetry). The input bits are taken in blocks of $2^t r+1$ and determine 2^t signal points for transmission. Of the $2^t r+1$ bits in each block, one bit determines whether any of the signal points should be an outer signal. If so, then t additional bits are used to determine in which of the 2^t intervals an outer signal will be sent, r-t bits determine which outer signal will be sent, and $r(2^t-1)$ bits, taken r at a time, are used to select the inner signals to be sent in the other 2^t-1 intervals. If none of the signal points is to be an outer signal, the remaining

$2^t r$ bits, taken r at a time, are used to select the inner signals to be sent in the 2^t intervals. The required signal-to-noise ratio for such a block coding system is about 3 x 2^{-t} dB greater than for an uncoded system in which a signal constellation having 2^r signals is used to send r bits per interval.

For example, referring to Fig. 9, to send 4 1/4 bits per interval, in 17-bit blocks each 4 intervals long (r=4, t=2), a signal constellation having 16 inner signals and 4 outer signals is used. One input bit determines whether any of the four output signal points will be an outer signal. If so, two further bits determine in which of the four intervals the outer signal will be sent, two further bits select that outer signal, and 12 bits, taken 4 at a time, select the three inner signals. If no outer signal is to be sent, the remaining 16 bits, taken 4 at a time, are used to select the four inner signals. The required signal-to-noise ratio to send 4 1/4 bits per interval is 7/8x10+1/8x26=12 (10.79 dB) or .79 dB more than to send 4 bits per interval in an uncoded system. (In this case, the outer signals might be moved to the axes to save a little more power, if maintenance of the grid is not required.)

In other embodiments, a system that sends N=r+f bits per modulation interval using a constellation having S signals can be extended to one sending $N'=r+f+2^{-t}$ bits per interval using a constellation having $(1+2^{-t})S$ signals. To the S original signals are added $2^{-t}S$ additional signals (assuming $2^{-t}S$ is an integer, and a multiple of 4 if quadrantal symmetry is to be maintained). These additional signals are further segmented into subgroups in the same proportion as in the unextended system; i.e., each subgroup is

2^{-t} as large as in the unextended system, and can therefore be specified by t fewer bits. For example, if in the unextended system there are $S=1.5x2^r$ signals divided into 2^r inner signals and 2^{r-1} outer signals, then in the extended system the $2^{-t}S$ additional points are divided into 2^{r-t} additional inner signals and 2^{r-t-1} additional outer signals.

Input bits are grouped into blocks of $2^t r+2^t f+1$ bits (assuming $2^t f$ is an integer) and used to determine 2^t signal points. One of these bits determines whether any of the additional $2^{-t}S$ signals is to be used; if not, the remaining $2^t(r+f)$ bits are used to determine the 2^t signal points using the unextended system; if so, then t bits are used to determine which interval uses an additional signal and the remaining $2^t(r+f)-t$ are used to determine which signals, using the unextended system, except that in the selected interval one of the additional signals is used and is selected with t fewer bits because the corresponding subgroups are all a factor of 2^{-t} smaller.

In this iterative fashion, systems that send N=r+f bits where f is any binary fraction of the form

$$f=2^{-t_1}+2^{-t_2}+\ldots+2^{-t_k}, \quad t_1<t_2<\ldots<t_k$$

can be built up, with signal constellations of

$$S=2^r(1+2^{-t_1})(1+2^{-t_2})\ldots(1+2^{-t_k})$$

signals (provided that $t_1+t_2+\ldots+t_k$ is less than or equal to r, or to r+2 if quadrantal symmetry is desired).

<u>4-Space Block Coded Systems</u>

In other embodiments, 4-space block coded modulation systems send an integer number (N) of bits per modulation interval using signal constellations of

the type of Fig. 7, or a half-integer number ($N=r+1/2$) bits per interval using constellations of the type of Fig. 1.

To send a half-integer number of bits per interval, $2r+1$ bits are grouped into a block which determines two signal points from a constellation having 2^{r+1} signals. The constellation is divided into two groups A and B of equal size, the two groups respectively containing alternate signals. For example, in Fig. 10 a constellation having 32 signals is divided into two 16-signal groups. As another example (not using a rectangular grid), Fig. 11 shows an 8-signal (8-phase) constellation divided into two 4-signal groups. The minimum squared distance between signals belonging to the same group ($d^2(A,A)=d^2(B,B)$) is at least twice as great as the minimum squared distance between signals belonging to different groups ($d^2(A,B)=d_o^2$). In Fig. 10 (and in any rectangular grid structure), $d^2(A,A)=2d^2(A,B)$; in Fig. 11 $d^2(A,A)=3.4d^2(A,B)$.

One bit from each block (the group bit) determines whether both signal points will be from group A or both from group B. The remaining 2^r bits taken r at a time determine which of the 2^r A points or B points is selected for each interval. The minimum squared distance between code words must be at least $2d_o^2$, because (a) between two code words that have different group bits (e.g., one group A code word and one group B code word), there must be a squared distance of at least $d^2(A,B)=d_o^2$ in each of the two intervals, while (b) between two code words with the same group bit, there must be a squared distance of at least $d^2(A,A)=d^2(B,B)\geqq 2d_o^2$ in one interval. However, the power used is only that required for

sending r+1 bits per interval uncoded with a minimum squared distance of d_o^2, which is about the same as for an uncoded system sending r bits per interval with a minimum squared distance of $2d_o^2$. Thus, 4-space coding allows sending r+1/2 bits per interval at about the same required signal-to-noise ratio as an uncoded system sending r bits per interval.

An embodiment using the Fig. 10 32-signal constellation to send 4 1/2 bits per modulation interval is as follows.

Referring to Fig. 12, in the signal selection logic, the first input bit (b_1) is used by group selector 40 to determine the groups ($group_1$ and $group_2$) from which both signal points will be selected, input bits b_2 through b_5 are used by signal point selector 42 to select signal $point_1$ from the named group, and input bits b_6 through b_9 to select signal $point_2$ from the named group.

The minimum squared distance between code words is 4, and the required signal-to-noise ratio $\overline{P}$ is 10, just as for the 16-signal uncoded Fig. 1 constellation. Thus 4 1/2 bits per interval can be sent with the 4-space coded system at the same $\overline{P}$ as 4 bits per interval uncoded.

At the receiver, the received code word (i.e., the two received signal points for each block) may be decoded by determining the most likely two A signal points to have been sent and the most likely two B signal points to have been sent (using 2-dimensional slicing in each modulation interval), and choosing the pair (i.e., the code word) which has the minimum squared distance to the received word.

In other embodiments, decoding is accomplished by making a tentative decision (in each modulation interval) of the most likely signal point to have been sent in that interval (using two-dimensional slicing with the two-dimensional coordinates being rotated 45° with respect to the signal constellation). If the tentative decisions in both intervals are of signal points in the same group, those tentative decisions become the final decision. Otherwise, the tentative decision having the least reliable coordinate is changed to the next nearest signal point (which will be from the other group), and the new tentative decision becomes the final decision.

In other embodiments, maximum likelihood decoding can be accomplished by applying the Viterbi algorithm. Referring to Fig. 13, the encoding process can be characterized by a finite-length 2-state trellis that determines the groups from which signal points are selected in two successive intervals in a manner analogous to Fig. 6. In this case Viterbi algorithm decoding is essentially equivalent to the first decoding method for 4-space codes described above.

Table IV shows the numbers of signals in the constellation and the required signal-to-noise ratios for sending half-integer numbers (between 3 1/2 and 7 1/2) of bits per interval using this 4-space coding system.

Table IV

N	S	$\overline{P}$	(db)
3 1/2	16	5	7.0
4 1/2	32	10	10.0
5 1/2	64	21	13.2
6 1/2	128	41	16.1
7 1/2	256	85	19.3

By dividing the groups of signals of the 4-space block coded system into subsets (in the manner previously described), quadrantal differential coding on a block basis can be used in the same manner previously described for the 8-space system.

To send an integral number N bits per interval using 4-space coding, the signal constellations for sending N+1/2 bits per interval described earlier in the section headed <u>Multidimensional Signal Structures for Sending a Non-Integral Number of Bits per Modulation Interval</u>, are used. In two dimensions, these constellations have $1.5x2^N$ signals, as shown in Fig. 7, for example. Because they are based on a rectangular grid, they may be divided into A and B groups of equal size ($1.5x2^{N-1}$ signals in each) by assigning alternate signals to the two groups. The minimum squared distance between signals in the same group is twice the minimum squared distance between signals in different groups. In coding, one input bit (the group bit) from a block of 2^N input bits determines whether A or B signals are to be used in both of two intervals. A second input bit determines whether an outer signal is to be used; if not, the remaining 2(N-1) bits (taken N-1 at a time) determine the two inner signals of the appropriate group; if so, a third bit determines which interval will contain an outer signal, N-2 bits select which outer signal of the appropriate group, and the remaining N-1 bits select the inner signal of the appropriate group in the other interval. For the same reasons as before, the minimum squared distance between code words is $2d_o^2$. The coding gain is about 1.5 dB. Any of the previously described decoding methods can be used.

To send 4 bits per interval, the 24-signal Fig. 7 constellation is rotated 45° and divided into A and B groups as well as into inner and outer signals, as shown in Fig. 14. Input bits are grouped into blocks of 8 bits each and each block determines two signal points. Referring to Fig. 15, in the signal selection logic, one bit (b_1), the group bit in each block, is used by group selector 50 to specify the group from which both signal points are to be selected, and a second bit (b_2) is used by outer point selector 52 to specify whether one of the signal points should be an outer signal. (At most one of the signal points for each block may be an outer signal.) If neither signal point is to be an outer signal, the remaining 6 (i.e., 2(N-1)) bits (b_3-b_8) taken three at a time are used by signal point selector 54 to determine which of the eight inner signals of the selected group is to be selected for each modulation interval. If one signal point is to be an outer signal, the third bit, b_3, determines which of the two signal points is to be an outer signal and the bits b_4-b_8 select the outer and inner signal as shown in Fig. 15.

The average power required to transmit inner signals is $\overline{P}_{inner} = 5$, and to transmit outer signals is $\overline{P}_{outer} = 13$. On average, half of the time the two signal points will both be inner signals, and the other half of the time one will be an inner signal and the other an outer signal. The average power required to send both signal points for a block is therefore $1/2(2 \times \overline{P}_{inner}) + 1/2(\overline{P}_{inner} + \overline{P}_{outer})$ or 14. The required signal-to-noise ratio is then 7 (8.45 dB) per interval compared with 10 (10 dB) required for sending uncoded signal points using the 16-point signal constellation of Fig. 1, a 1.55 dB coding gain.

Theoretically, as shown by Conway and Sloane, the best known 4-space code for sending 4 bits per interval has a required signal-to-noise ratio of 6.75 (8.29 dB). The present embodiment which has a required signal-to-noise ratio of 7 (8.45 db) provides a simpler coding system at a cost of only 0.16 db compared with the best known code for the same parameters. Further, because the code word set has quadrantal symmetry, quadrantal differential coding may be used (as described above) to obtain immunity to phase rotations of multiples of 90°.

Table V shows the numbers of constellation signals S and the required signal-to-noise ratios $\overline{P}$ for sending integer numbers N (from 4 to 7) of bits per interval using this system.

Table V

N	S	$\overline{P}$	(dB)
4	24	7	8.45
5	48	14.25	11.54
6	96	28.5	14.5
7	192	58	17.6

It can be seen that a coding gain of about 1.5 dB is obtained for all values of N with a signal constellation only 1.5 times the size of that needed for an uncoded system.

Further 8-Space Block Coded Systems

The 4-dimensional signal structures with 2-dimensional constellations like those of Fig. 7 can also be used in 8-space block coded systems by dividing the signals into subsets A_0, A_1, B_0, B_1 in the manner described above. The resulting systems send half-integer numbers of bits per modulation interval with about 3.0 dB coding gain over the uncoded 4-dimensional signal structures. Table VI is an

extension of Table II showing some of these half-integer systems in relation to the integer systems previously described.

Table VI

N	S	$\overline{P}$	(dB)
3	16	2.5	4.00
3 1/2	24	3.5	5.45
4	32	5.0	7.00
4 1/2	48	7.125	8.54
5	64	10.5	10.22
5 1/2	96	14.125	11.51

16-Space Block Coded Systems

In other embodiments, 16-space block coded modulation systems are used to send an integer number N of bits per modulation interval using Fig. 7-type constellations, or half-integer numbers using Fig. 1-type constellations.

The constellation must be further partitioned into eight equal-sized classes A_{00}, A_{01}, A_{10}, A_{11}, B_{00}, B_{01}, B_{10}, B_{11}, after first being divided into equal-sized groups A and B and dividing each group into equal-sized subsets A_0, A_1, and B_0, B_1. At each step the division is made by taking alternate signals. Furthermore, subsets are divided into two classes with opposite subscripts; e.g., A_0 is divided into A_{00} and A_{11}, A_1 into A_{01} and A_{10}, and so forth. The result is that if the minimum squared distance between constellation signals is $d_o^2=d^2(A,B)$, the minimum squared distance between signals in classes in the same group is $2d_o^2$ times the Hamming distance (d_H) between their subscripts, e.g., $d^2(A_{00}, A_{01})=2d_o^2$; $d^2(A_{00},A_{11})=4d_o^2$; and the minimum squared distance between signals in the same class is $8d_o^2$. Fig. 16 shows the 48-signal constellation of Fig. 7 rotated 45° and partitioned in this way.

In coding, one input bit of a block of 8N input bits determines the group (A or B) from which all 8 signal points will be selected. Eleven input bits are coded in a conventional (16, 11) Hamming code (with minimum Hamming distance equal to 4; see, e.g., Berlekamp, cited above) to give 16 coded bits, which are taken two at a time to determine the class subscripts for each signal point. The remaining 8N-12 input bits select which signal points are used; if N is an integer, by using the 4-dimensional system signal constellations of $1.5\times2^{N-1}$ signals and taking the bits 2N-3 at a time to determine pairs of signal points as described above; if N is a half-integer (N=r+1/2), then by using 2-dimensional signal constellations of 2^{r+2} signals and taking bits N-3/2=r-1 at a time to determine signal points from the selected classes.

The minimum squared distance of such 16-space block code words is $8d_o^2$. Two code words with different group bits must differ by at least d_o^2 in all 8 signal points. Two code words with the same group bit but different Hamming coded bits must differ in at least four class subscripts, so their squared distance is at least $4\times2d_o^2$. Two code words with the same group bit and Hamming coded bits have all signal points in the same classes, but must have in at least one interval two different signal points in the same class, which must therefore have squared distance $8d_o^2$. The minimum squared distance between code words is therefore 8 times (or 9 dB) better than that of the uncoded constellation, but that constellation could without coding send 1 1/2 more bits per interval than it can with coding, which is equivalent to about 4.5 dB, so the net.coding gain is about 4.5 dB.

An embodiment using the Fig. 16 48-signal constellation to send 4 bits per interval is as follows.

Referring to Fig. 17, 32 input bits are used to select the 8 output signal points as follows. One bit is used by group selector 70 to the determine the group (A or B) from which all signal points will be selected. Eleven bits are coded by Hamming coder 72 using a conventional (16,11) extended Hamming code (with Hamming distance d_H=4) into 16 coded bits. Taken in pairs, the 16 coded bits are used by class selector 74 to specify the subscripts of the classes from which the eight output signal points are respectively selected. The remaining 20 bits (taken 5 at a time) are used by signal point selector 76 to select pairs of signal points of appropriate classes, where each class comprises 6 signals, 4 inner and 2 outer, and selection between inner and outer signals is made in the same manner as previously described.

The required signal-to-noise ratio is 3.5625 or 5.5dB, a 4.5dB gain over the uncoded N=4 constellation of Fig. 1.

Table VII shows the number of signals S in the constellation and the required signal-to-noise ratio $\overline{P}$ for sending integer and half-integer numbers N (between 2 1/2 and 6 1/2) of bits per interval in blocks of 8 intervals each, using the 16-space coding system described above and the signal constellations of Figs. 1 and 7.

Table VII

N	S	$\overline{P}$	(dB)
2 1/2	16	1.25	1.0
3	24	1.75	2.5
3 1/2	32	2.5	4.0
4	48	3.56	5.5
4 1/2	64	5.25	7.2
5	96	7.13	8.5
5 1/2	128	8.25	10.1
6	192	14.5	11.6
6 1/2	256	21.25	13.3

In the signal constellations for the 16-space block coding systems, each signal is a member of a quadruplet of signals at the same radius from the origin and separated by intervals of 90° about the origin. (For example, the points marked 60 and 62 in Fig. 16 form such quadruplets.) Because all such quadruplets are of one or the other forms shown in Fig. 18, quadrantal differential coding (in the manner previously described) can be used with the group bit and an appropriate one of the class subscript bits (in this case the second for A and the first for B, inverted) in place of the group and subset bits used in 8-space coding, but otherwise as previously described.

24-Space Block Coded Systems

In other embodiments, 24-space block coded systems can be used to send an integer number of bits per interval using Fig. 1-type constellations, or a half-integer number of bits per interval using Fig. 7-type constellations.

To send N bits per interval, N an integer, a constellation of 2^{N+2} signals on a rectangular grid (such as one of those of Fig. 1) is used. The constellation is divided into 16 subclasses A_{ijk}, B_{ijk} (i,j,k=0 or 1) by further subdivision of the classes A_{ij} and B_{ij} defined in the previous section by taking alternate signals in accordance with a pattern defined below. The minimum squared distance between signals in different groups is then d_o^2, between signals in the same group but different subsets is $2d_o^2$, in the same subset but different classes is $4d_o^2$, in the same class but different subclasses is $8d_o^2$, and finally between signals in the same subclass is $16d_o^2$.

Referring to Fig. 19, the 64-signal constellation of Fig. 1 has been partitioned into 16 4-point subclasses (for use with a 24-space block coding system). In general, a partition of any rectangular-grid constellation into subclasses using the pattern of subscripts shown in Fig. 19 may be used.

In coding, one input bit of a block of 12N bits determines the group (A or B) from which all 12 signal points will be selected. Twelve input bits are coded in the (24, 12) Golay code (which has minimum Hamming distance 8; see, e.g., Berlekamp, cited above) to give 24 coded bits, which are taken two at a time to give the (i, j) subscripts of the subclass from which each signal point is to be selected. Eleven input bits determine the k subscripts for the first 11 points, and the 12th k subscript is chosen so that the 12 k subscripts have even parity (an even number are equal to 1) for A code words and odd parity for B code words. The remaining 12N-24 bits are taken N-2 at a time to select one of the 2^{N-2} signals of the selected subclass for each of the 12 intervals.

With this construction, the minimum squared distance between code words is $16d_o^2$, where d_o^2 is the minimum squared distance between signals in the constellation. This is 12 dB better than the uncoded constellation having 2^{N+2} signals, but that constellation without coding could send 2 bits per interval more than with this coding, which is equivalent to about 6 dB, so the net coding gain is about 6 dB.

One embodiment, using the Fig. 19 constellation to send 4 bits per interval, is as follows. Referring to Fig. 20, 48 input bits are used to select 12 output signal points as shown. One bit is used by group selector 80 to determine the group (A or B) from which

all signal points will be selected. Twelve bits are coded by Golay coder 82 using a conventional (24, 12) Golay code (with Hamming distance $d_H=8$) into 24 coded bits. Taken in pairs, the 24 coded bits are used by class selector 84 to specify the (i, j) subscripts of the 12 signals to be selected. Eleven bits and the group bit are used by subclass selector 86 to specify the k subscripts of the 12 points, with 11 subscripts equal to the corresponding bits, and the 12th subscript a parity check on these 11 bits plus the group bit. The remaining 24 bits, taken two at a time, are used by signal point selector 88 to select one of the four signals in each selected subclass.

The required signal-to-noise ratio with this system is 2.625 or 4.2dB, 5.8dB better than with the uncoded N=4 constellation of Fig. 1.

Table VIII shows the numbers of signals in the constellation and the required signal-to-noise ratios $\overline{P}$ for sending integer numbers N (between 2 and 6) of bits per interval in blocks of 12 intervals, using the 24-space coding described above and the signal constellations of Fig. 1. (The constellations of Fig. 7 could be used to send half-integer numbers of bits per interval with similar coding gain.)

Table VIII

N	S	$\overline{P}$	(dB)
2	16	.625	-2.0
3	32	1.255	1.0
4	64	2.625	4.2
5	128	5.125	7.1
6	256	10.625	10.3

In the signal constellations for these 24-space block coded modulation systems, each signal is a member of a quadruplet of signals of the same radius and

separated by phase intervals of 90°. (For example, the signals marked 90, 92, 94, and 96 in Fig. 19 form such quadruplets, respectively.) All such quadruplets are of one of the four types shown in Fig. 21, and therefore quadrantal differential coding (in the manner previously described) can be used with the group bit and an appropriate subscript bit (in this case the first for A; and the second for B, inverted) in place of the group and subset bits used in 8-space coding, but otherwise as previously described.

Block Coded Systems of Other Numbers of Dimensions

In other embodiments, block coded systems in 2m-space can be derived from the 4-space and 8-space block coded systems.

The 4-space system can be extended to a 2m-space system that sends N-1/m bits per interval or mN-1 bits per block (for any m greater than or equal to 2) if N is an integer, or for any even m greater than or equal to 2 if N=r+1/2 is a half-integer number. In the former case the signal constellation of size 2^N is used, and in the latter the constellation of $1.5x2^r$ signals divided into inner and outer signals as previously described. In either case the constellation is divided into A and B groups as previously described. The encoder is arranged so that in the first m-1 intervals, either A or B signals may be sent, but in the last interval, the group must be constrained to satisfy a parity condition (e.g., that the total number of A signals be even). For example, the groups may be selected by a two-state trellis encoder operating according to the m-interval trellis of Fig. 22. Otherwise selection of signals within the groups is as previously described. Decoding may be by the Viterbi algorithm.

The minimum squared distance between code words remains $2d_o^2$ in this system because of the parity condition and the fact that the minimum squared distance between signals within the same group is at least twice the minimum squared distance between signals in different groups. However, the number of bits transmitted per interval is now only 1/m less than in an uncoded system using the same constellation, so that the coding gain is approximately 3(1-1/m) dB, which approaches 3 dB for m large. For example, for m=4, an 8-space coded modulation system with coding gain of about 2.25 dB is obtained.

The 8-space system can be extended to a 2m-space system that sends N bits per interval for any m greater than or equal to 4 and any integer N, using a signal constellation of 2^{N+1} signals divided into 4 equal-sized subsets as previously described. The encoder is arranged so that all m points are chosen from the same group, the first m-1 points being from either subset of that group, and the final point being chosen from a subset selected to satisfy a parity condition (e.g., that the total number of 1 subscripts be even). For example, the subsets may be selected by a 4-state trellis encoder that operates according to the m-interval trellis of Fig. 23. Otherwise selection of signals within subsets is as previously described. Decoding may be done using the Viterbi algorithm using the Fig. 23 trellis.

The minimum squared distance between code words remains $4d_o^2$ in this system because m is greater than or equal to 4, there is a parity condition on subsets combined with the between-subset distance of $2d_o^2$, and there is a within-subset distance of $4d_o^2$. The number of bits per interval is N for any value of m, so the coding gain remains about 3 dB.

However, for m greater than 4 the number of near neighbors of any code word is reduced since code words of the opposite group have distance at least md_0^2 and are thus no longer near neighbors, reducing the error event probability coefficient by about a factor of 2.

Other Embodiments

Other embodiments are within the following claims. For instance, the input bits can undergo any transformation which does not result in loss of information (such as scrambling, permutation, or binary linear transformations) before being encoded. The digital data to be sent can be in characters other than bits. The output signal points or indeed their coordinates can be arbitrarily permuted without loss of coding gain. The coordinates can be used individually in one-dimensional modulation systems (such as single side-band or vestigial-sideband modulation systems), or in groups of appropriate dimension in multidimensional modulation systems. Other signal constellations (including ones not based on rectangular grids) can be used provided that their signals can be partitioned into equal-sized groups (subsets, classes, subclasses) having at least approximately the desired distance properties described earlier. Any 2m-dimensional orthogonal linear transformation of output signal points may be made without affecting relative distance properties in 2m-space. Quadrantal differential coding can be done using any method such that 90° rotations of code words change only quadrantal phase bits, and such phase bits are encoded differentially (mod 4) with reference to any previously transmitted quadrantal phase bits. Multidimensional signal constellations for any binary fractional numbers of bits per interval as described herein, may be used with the block coded modulation systems provided that parameters are consistent.

CLAIMS

1. A method for sending digital data over a band-limited channel comprising selecting a code word corresponding to a plurality of modulation signal points drawn from a constellation of available signals, the code word being selected from an available set of said code words on the basis of a block of said digital data, and the code words belonging to said available set for said block being independent of said signal points corresponding to the code word selected for any other said block, characterised in that said constellation comprises groups having equal numbers of said signals, and in that the group from which at least one said signal point for said block is drawn depends on the group from which at least one other said signal point for said block is drawn.

2. A method according to Claim 1, further characterised in that said signal points are two-dimensional, said code words are 2m-dimensional, m being the number of said signal points corresponding to each said code word, and each said block of said digital data comprising mN bits.

3. A method according to Claims 1 or 2, further characterised in that said signals of said constellation are arranged on a rectangular grid.

4. A method according to Claim 1, further characterised in that said signals are arranged in said constellation so that the minimum squared distance between two said signals belonging to the same said group is greater than between two said signals belonging to different said groups.

5. A method according to Claim 4, further characterised in that the minimum squared distance between two said signal points belonging to the same said group is twice the minimum squared distance between two said signal points belonging to different said groups.

6. A method according to any preceding claim, further characterised in that each said block comprises a plurality of bits, and in that the said groups from which said signal points for said block are drawn are determined by a single said bit of said block.

7. A method according to Claim 2, further characterised in that m-1 said signal points for each said block are drawn from among all said signals in said constellation, and the one remaining said signal point for said block is drawn from a group that depends on the groups from which said m-1 signal points are drawn.

8. A method according to Claim 7, further characterised in that said group from which said one remaining signal point is drawn depends on a single-parity-check code based on at least one of said bits.

9. A method according to Claim 9, further characterised in that said group from which said one remaining signal point is drawn is determined on the basis of a state-transition trellis for said block.

10. A method according to Claim 2 or any of Claims 3 to 9 when appendent thereto, further characterised in that N is an integer, m is an even integer no smaller than 2, and said constellation comprises $1.5x2^{N}$ signals.

11. A method according to Claim 2 or any of Claims 3 to 9 when appendent thereto, further characterised in that N=r+1/2, r being an integer, m is no smaller than 2, and said constellation comprises 2^{r+1} signals.

12. A method according to any of Claims 2 to 11, further characterised in that m is 2.

13. A method according to any of Claims 1 to 3, further characterised in that said groups each comprise subsets having equal numbers of said signals, and in that the subset from which at least one said signal point for said block is drawn depends on the subset from which at least one other said signal point for said block is drawn.

14. A method according to Claim 13, further characterised in that said signals are arranged in said constel-

lation so that the minimum squared distance between two signals belonging to one said subset is greater than between two said signals belonging to different said subsets within the same said group.

15. A method according to Claim 14, further characterised in that the minimum squared distance between two said signal points belonging to the same said subset is twice the minimum squared distance between two said signal points belonging to different said subsets.

16. A method according to any of Claims 13 to 15, further characterised in that each said block comprises a plurality of bits, and the subsets from which said signal points for said block are drawn are determined based on a plurality of bits representing less than all of said digital data of said block.

17. A method according to both Claims 2 and 13, further characterised in that one said signal point for each said block is drawn from among all said signals in said constellation, m-2 of said signal points are drawn from groups that depend on the group from which said one signal point is drawn, and the single remaining said signal point for said block is drawn from a subset that depends on the subsets from which said one signal point and said m-2 signal points are drawn.

18. A method according to Claim 17, further characterised in that said subset from which said single remaining signal is drawn depends on a single-parity-check code based on at least one of said bits.

19. A method according to Claim 17, further characterised in that said subsets from which said signal points for said block are drawn are determined on the basis of a state-transition trellis for said block.

20. A method according to Claim 17, further characterised in that said subsets and said groups from which said signal points for said block are drawn are determined based on a Hamming code applied to at least one of said bits.

21. A method according to any of Claims 13 to 20 when appendent to Claim 2, further characterised in that N is an integer, m is at least 4, and said constellation comprises 2^{N+1} said signals.

22. A method according to any of Claims 13 to 20 when appendent to Claim 2, further characterised in that N=r+1/2, r being an integer, m is an even integer no smaller than 4, and said constellation comprises $1.5x2^{r+1}$ signals.

23. A method according to any of Claims 13 to 22 when appendent to Claim 2, further characterised in that m is 4.

24. A method according to Claim 13, further characterised in that said subsets each comprise classes having equal numbers of said signals, and in that said class from which at least one signal point for said block is drawn depends on the class from which at least one other said signal point for said block is drawn.

25. A method according to Claim 24, further characterised in that said signals are arranged in said constellation so that the minimum squared distance between two signals belonging to one said class is greater than between two said signals belonging to different said classes within the same said subset.

26. A method according to Claim 25, further characterised in that the minimum squared distance between two said signal points belonging to the same said class is twice the minimum squared distance between two said signal points belonging to different said classes.

27. A method according to any of Claims 24 to 26, further characterised in that each said block comprises a plurality of bits, and the classes from which said signal points for said block are drawn are determined based on a plurality of bits representing less than all of said digital data of said block.

28. A method according to any of Claims 24 to 27, when appendent to Claim 7, further characterised in that m is 8.

29. A method according to Claim 28, further characterised in that N is an integer and said constellation comprises $1.5x2^{N+1}$ said signals.

30. A method according to Claim 28, further characterised in that N=r+1/2, r being an integer, and said constellation comprises 2^{r+2} signals.

31. A method according to Claim 24, further characterised in that said classes from which said signal points for said block are drawn are determined based on a Hamming code applied to at least one of said bits.

32. A method according to Claim 24, further characterised in that said classes each comprise subclasses having equal numbers of said signals, and in that said subclass from which at least one signal point for said block is drawn depends on the subclass from which at least one other said signal point for said block is drawn.

33. A method according to Claim 32, further characterised in that said signals are arranged in said constellation so that the minimum squared distance between two signals belonging to one said subclass is greater than between two said signals belonging to different said subclasses within the same said class.

34. A method according to Claim 33, further characterised in that the minimum squared distance between two said signal points belonging to the same said subclass is twice the minimum squared distance between two said signal points belonging to different said subclasses.

35. A method according to any of Claims 32 to 34, further characterised in that each said block comprises a plurality of bits, and the subclasses from which said signal points for said block are drawn are determined based on a plurality of bits representing less than all of said digital data of said block.

36. A method according to any of Claims 32 to 35, when appendent to Claim 2, further characterised in that m is 12.

37. A method according to Claim 36, further characterised in that N is an integer and said constellation comprises 2^{N+2} said signals.

38. A method according to Claim 33, further characterised in that said subclasses from which said signal points for said block are drawn are determined based on a Golay code applied to at least one of said bits.

39. A method according to Claim 2 or any of Claims 3 to 38 (other than Claims 10, 22 or 30) when appendent to Claim 2, further characterised in that N is 4.

40. A method according to Claims 10, (22 or 30,) or any Claim appendent thereto, further characterised in that N is 4½.

41. A method according to any preceding Claim, further characterised in that said constellation comprises quadruplets each having four said signals, the signals belonging to each said quadruplet being located at the same distance from the origin but separated by 90^{o} intervals about said origin, said digital data comprises bits, and at least two of said bits are quadrantally differentially encoded.

42. A method according to Claim 41, further characterised in that said groups each comprise subsets having equal numbers of said signals and said four signals belonging to each quadruplet are drawn from four different said subsets.

43. A method according to any preceding claim, further characterised in that decoding is conducted by deciding which code word was sent based on maximum likelihood sequence estimation in accordance with the Viterbi algorithm.

44. A method according to any of Claims 1 to 40, further characterised in that all said signal points for each said code word are drawn from the same said group.

45. A method according to Claim 44, further characterised in that decoding is conducted by deciding which said code word was sent by first making a separate tentative decision based on code words whose signal points are drawn from each of said groups, and thereafter making a final

decision based on said separate tentative decisions.

46. A method according to any of claims 1 to 40, further characterised in that said digital data comprises bits and the selection of said signal points for a block depends on a single-parity-check code based on at least one of said bits.

47. A method according to Claim 46, further characterised in that decoding is conducted by deciding which said code word was sent by first making a tentative decision as to each signal point in said code word without regard to whether said parity check is satisfied, and accepting said tentative decisions as the final decision, either without changes if said parity check is satisfied, or after changing the least reliable one of said tentative decisions if said parity check is not satisfied.

48. A method for sending a block of digital data bits over a band-limited channel using a plurality of modulation signal points drawn from a two-dimensional constellation of available signals, characterised in that said constellation comprises a plurality of inner signals, and a plurality of outer signals located farther from the origin than said inner signals, in that one bit of said digital data determines whether any of said plurality of signal points will be drawn from said outer signals, and in that if an outer signal will be drawn, at least one other bit of said digital data determines which of said plurality of signal points will be an outer signal point.

49. A method according to Claim 48, further characterised in that there are 2^t said signal points, said digital data comprises at least t+1 bits, there are S said inner signals and $2^{-t}S$ said outer signals, and t said bits determine which of said plurality of signal points will be an outer signal point.

50. A method according to Claims 48 or 49, further characterised in that said signals of said constellation are arranged on a rectangular grid.

51. A method according to Claims 48 to 50, further characterised in that said constellation comprises quadruplets each having four said signals, the signals belonging to each said quadruplet being located at the same distance from the origin but separated by 90^o intervals about the origin of a signal plane, and at least two of said bits are quadrantally differentially encoded.

52. A method according to Claim 49, further characterised in that t=1.

53. A method according to Claim 52, further characterised in that said block comprises 2N+1 said bits and S is 2^N.

54. A method according to Claim 53, further characterised in that N of said bits determine which inner signal is drawn.

55. A method according to Claim 49, further characterised in that t is 2.

56. A method according to Claim 55, further characterised in that said block comprises 4N+1 bits and S is 2^N, N being an integer.

57. A method according to Claim 55, further characterised in that said block comprises 4N+3 bits and S is $1.5x2^N$, N being an integer.

58. A method according to Claim 56, further characterised in that N-2 of said bits determine which said outer signal is drawn and in that N of said bits determine which inner signal is drawn.

59. A method according to any of Claims 53, 54, 56, 57 or 58, further characterised in that N is 4.

60. Apparatus for sending digital data over a band-limited channel, characterised in that said apparatus comprises a code word selector for selecting code words corresponding to a plurality of modulation signal points drawn from a constellation of available signals, said code word selector being adapted to select a said code word from an available set of said code words on the basis

of a block of said digital data, the code words belonging to said available set for said block being independent of said signal points corresponding to the code word selected for any other said block, and said constellation comprising groups having equal numbers of said signals, and in that said apparatus includes means for determining the group from which at least one said signal point for said block is drawn in dependence upon the group from which at least one other said signal point for said block is drawn.

61. Apparatus according to Claim 60, wherein said groups each comprise subsets having equal numbers of said signals, said apparatus being further characterised in that it comprises subset selector means adapted to select the subset from which at least one said signal point for said block is drawn in dependence upon the subset from which at least one other said signal point for said block is drawn.

62. Apparatus according to Claim 61, further characterised in that said subset selector means is adapted to determine said subsets from which said signal points for said block are drawn on the basis of a state-transition trellis for said block.

63. Apparatus according to Claim 61, further characterised in that said subset selector means is adapted to determine the subsets and the groups from which said signal points for said block are drawn based upon a Hamming code applied to at least one of said bits.

64. Apparatus according to Claim 61, wherein said subsets each comprise classes having equal numbers of said signals, said apparatus being further characterised in that it includes class selector means adapted to select said class from which at least one signal point for said block is drawn in dependence upon the class from which at least one other said signal point for said block is drawn.

65. Apparatus according to Claim 64, further characterised in that said class selector means is adapted to determine said classes from which said signal points for said block are drawn based upon a Hamming code applied to at least one of said bits.

66. Apparatus according to Claim 64, wherein said classes each comprise subclasses having equal numbers of said signals, said apparatus being further characterised in that it includes subclass selector means adapted to select said subclass from which at least one signal point for said block is drawn in dependence upon the subclass from which at least one other said signal point for said block is drawn.

67. Apparatus according to Claim 66, wherein said subclass selector means is adapted to determine said subclasses from which said signal points for said block are drawn based upon a Golay code applied to at least one of said bits.

68. Apparatus according to any of Claims 60 to 67, characterised in that it further comprises means for selecting at least one said signal point of said block on the basis of less than all said digital data in said block.

69. Decoding apparatus characterised in that it is adapted to operate in conjunction with apparatus according to any of Claims 60 to 68 to decode digital data sent over a band-limited channel, and in that said decoder includes decider means adapted to decide which code word was sent, said decider means being adapted to operate on a maximum likelihood sequence estimation in accordance with the Viterbi algorithm.

70. Decoding apparatus characterised in that it is adapted to operate in conjunction with apparatus according to any of Claims 60 to 68 to decode digital data sent over a band-limited channel, and in that said decoder includes means adapted to make tentative decisions about which signal point was sent prior to

final decoding of all of the received signal points for a block, and in that said decoder includes adaptive control circuitry arranged to be responsive to said tentative decision.

71. Decoding apparatus characterised in that it is adapted to operate in conjunction with apparatus according to any of Claims 60 to 68 to decode digital data sent over a band-limited channel, and in that said apparatus includes means adapted to decide which said code word was sent, all said signal points for each said code word being drawn from the same said group, by first making a separate tentative decision based on code words whose signal points are drawn from each of said groups, and thereafter making a final decision based on said separate tentative decisions.

72. Decoding apparatus characterised in that it is adapted to operate in conjunction with apparatus according to any of Claims 60 to 68 to decode digital data sent over a band-limited channel, and in that said apparatus includes decider means adapted to decide which said code word was sent, said digital data comprising bits and the selection of said signal points for a block depending on a single-parity-check code based on at least one of said bits, which decider means includes first means for making a tentative decision as to each signal point in said code word without regard to whether said parity check is satisfied, and second means responsive to said parity check and adapted to accept said tentative decision as the final decision either without change if said parity check is satisfied or after changing the least reliable one of said tentative decisions if said parity check is not satisfied.

73. Apparatus for sending a block of digital data bits over a band-limited channel, said apparatus being characterised in that it comprises means for drawing a plurality of modulation signal points from a two-

dimensional constellation of available signals which comprises a plurality of inner signals and a plurality of outer signals located further from the origin that said inner signals, and in that said apparatus comprises means responsive to one bit of said digital data for determining whether any of said plurality of signal points will be drawn from said outer signals, and in that said apparatus includes means for determining on the basis of at least one other bit of said digital data, if an outer signal will be drawn, which of said plurality of signal points will be an outer signal point.

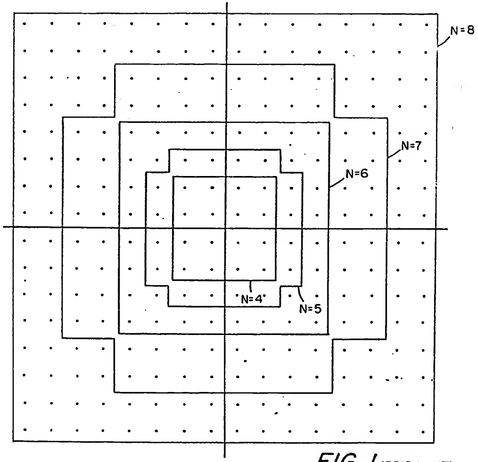


FIG. 1 (PRIOR ART)

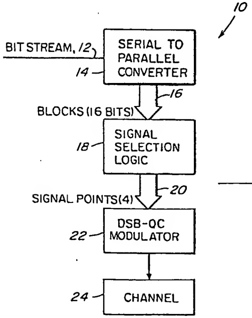


FIG. 2

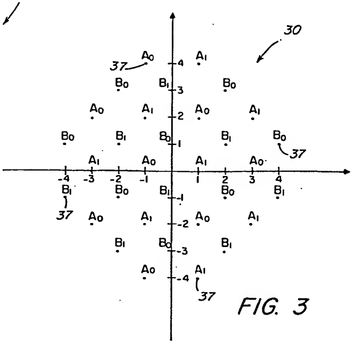


FIG. 3

0122805

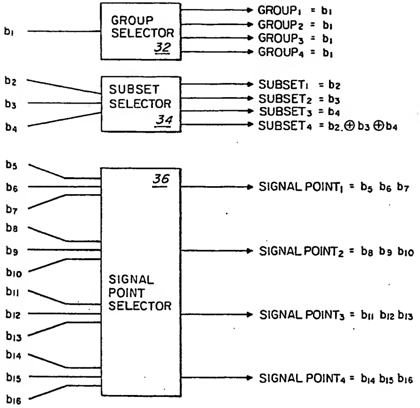


FIG. 4

SUBSET BIT	GROUP BIT	SIGNAL SUBSET
0	0	A_0
0	1	B_0
1	0	A_1
1	1	B_1

FIG. 5

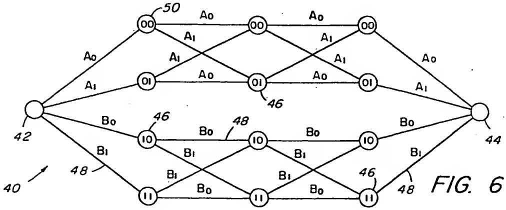


FIG. 6

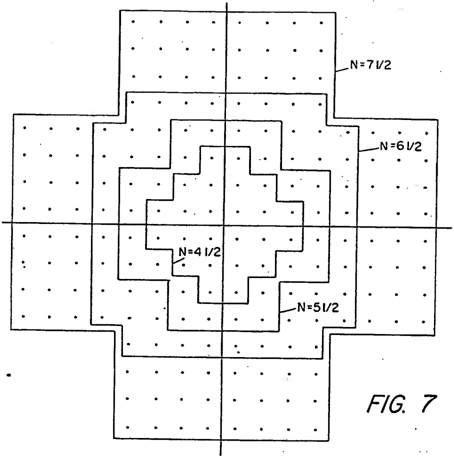
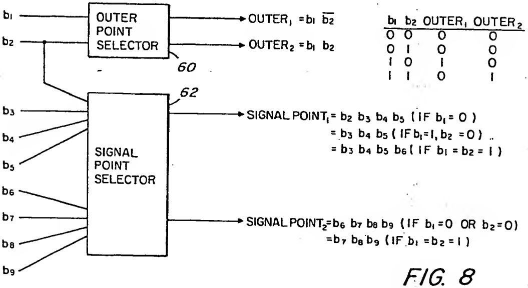


FIG. 7



b_1	b_2	OUTER$_1$	OUTER$_2$
0	0	0	0
0	1	0	0
1	0	1	0
1	1	0	1

FIG. 8

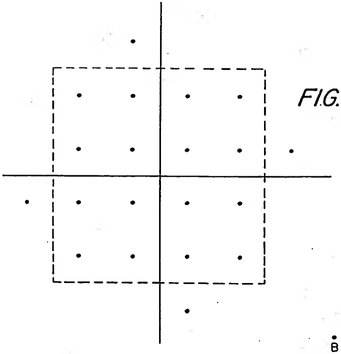


FIG. 9

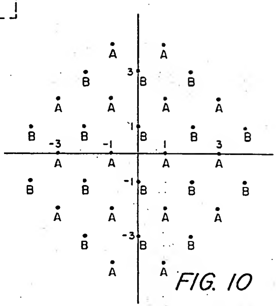


FIG. 10

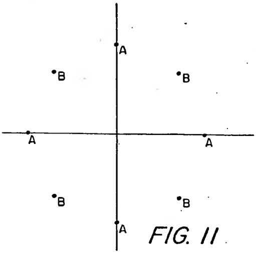


FIG. 11

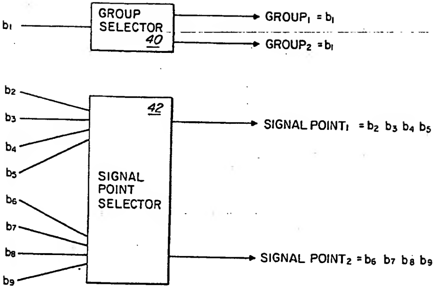


FIG. 12

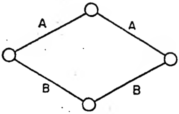


FIG. 13

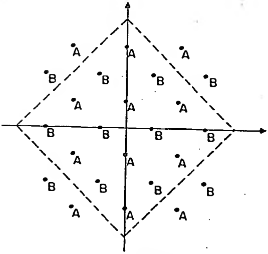


FIG. 14

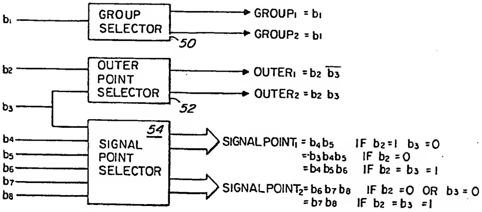


FIG. 15

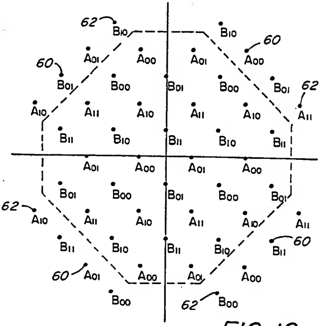
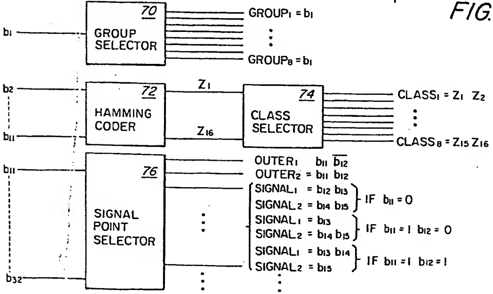


FIG. 16

FIG. 17



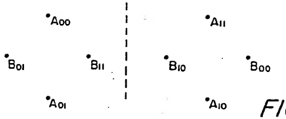


FIG. 18

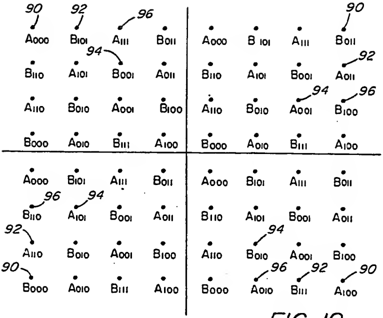


FIG. 19

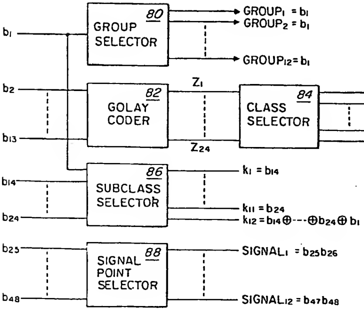


FIG. 20

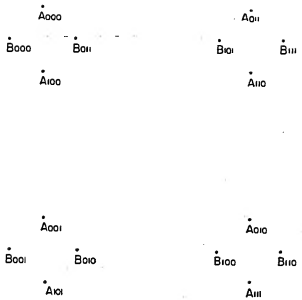


FIG. 21

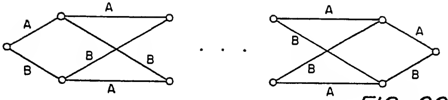


FIG. 22

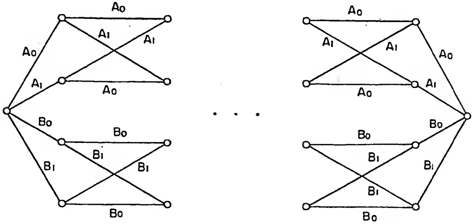


FIG. 23

METHOD AND EQUIPMENT FOR TRANSMITTING VIDEO SIGNAL

Patent Number: JP63180280
Publication date: 1988-07-25
Inventor(s): FURUHATA TAKASHI
Applicant(s):: HITACHI LTD
Requested Patent: JP63180280
Application Number: JP19870011399 19870122
Priority Number(s):
IPC Classification: H04N7/08 ; H04J1/00
EC Classification:
Equivalents: JP2528108B2

Abstract

PURPOSE:To transmit the video signals of two channels in the band for one channel by mutually frequency-multiplexing the video signals of a first channel and a second channel in the band for one channel.

CONSTITUTION:The video signals V1 and V2 of the first and the second channels are supplied to terminals 1 and 2 in a synchronized phase relation. The signal V2 is supplied to a switching circuit 20 and a phase inversion circuit 10, and a phase inverted output to the circuit 20. It is switched in the horizontal scanning line unit of the signal and the output is supplied to a synthesis circuit 30. The sum component of the signals V1 and V2 is outputted in the first signal block of the circuit 30, and the difference component of the signals V1 and V2 is outputted in a second signal block. The signals for two channels are frequency-multiplexed in the band for one channel. Thus, the signals for two channels can be transmitted in the band for one channel.

⑲日本国特許庁（JP）　⑪特許出願公開

⑫公開特許公報（A）　昭63−180280

⑤Int.Cl.⁴　識別記号　庁内整理番号　④公開　昭和63年(1988)7月25日

H 04 N 7/08　Z−7060−5C
H 04 J 1/00　8226−5K

審査請求　未請求　発明の数 3　（全11頁）

⑤発明の名称　映像信号の伝送方法及びその装置

⑪特　願　昭62−11399
⑫出　願　昭62(1987)1月22日

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⑭代理人　弁理士　並木　昭夫

明　細　書

1. 発明の名称

映像信号の伝送方法及びその装置

2. 特許請求の範囲

1. 伝送すべき第1チャンネルの映像信号と第2チャンネルの映像信号とを、各々の第1の信号ブロックでは該第1チャンネルの映像信号と第2チャンネルの映像信号との和に相当する位相関係で周波数多重し、前記第1の信号ブロック以外の第2の信号ブロックでは前記第1チャンネルの映像信号と第2チャンネルの映像信号との差に相当する位相関係で周波数多重して伝送するようにしたことを特徴とする映像信号の伝送方法。

2. 特許請求の範囲第1項に記載の伝送方法において、前記第1チャンネルの映像信号における第1の信号ブロックと第2の信号ブロックとの位相関係、及び前記第2チャンネルの映像信号における第1の信号ブロックと第2の信号ブロックとの位相関係は、それぞれ、各々の映像信号のフィールド内の時間的に隣接するライン同士か、あるいは隣接するフィールド間あるいはフレーム間の空間的に隣接するライン同士の関係に有ることを特徴とする映像信号の伝送方法。

3. 特許請求の範囲第1項に記載の伝送方法において、前記第1チャンネルの映像信号と第2チャンネルの映像信号は、共に、輝度信号と色差信号が時分割多重して成ることを特徴とする映像信号の伝送方法。

4. 特許請求の範囲第1項に記載の伝送方法において、前記第1チャンネルの映像信号は、互いに視差の異なる2つの立体映像情報に基づく2つの映像信号のうちの一方の映像信号から成り、前記第2チャンネルの映像信号は、他方の映像信号から成ることを特徴とする映像信号の伝送方法。

5. 第1チャンネルの映像信号及び第2チャンネルの映像信号を伝送するための伝送装置において、前記第1チャンネルの映像信号を入力

し、該第1チャンネルの映像信号における2つの色度信号のうちの少なくとも一方の色度信号と輝度信号とを時分割多重して出力する第1の時分割多重手段と、前記第2チャンネルの映像信号を入力し、該第2チャンネルの映像信号における2つの色度信号のうちの少なくとも一方の色度信号と輝度信号とを時分割多重して出力する第2の時分割多重手段と、該第1及び第2の時分割多重手段からの各出力信号を入力し、各々の信号におけるフィールド内の時間的に隣接するライン同士、あるいは隣接するフィールド間あるいはフレーム間の空間的に隣接するライン同士のうち、一方のラインにおいては、前記第1及び第2の時分割多重手段からの各出力信号の和に相当する位相関係でその両者を周波数多重し、もう一方のラインにおいては、前記第1及び第2の時分割多重手段からの各出力信号の差に相当する位相関係でその両者を周波数多重する周波数多重手段と、から成り、該周波数多重手段によって多重された信号を伝送するようにしたことを特徴とする映像信号の伝送装置。

6. 特許請求の範囲第5項に記載の伝送装置において、第1の映像信号を入力し、各ライン毎に該第1の映像信号の時間軸をN倍に伸長し、その後、フィールド内の時間的に隣接するライン同士、あるいは隣接するフィールド間あるいはフレーム間の空間的に隣接するライン同士のうち、一方のラインに相当する映像信号を第1出力として出力し、他方のラインに相当する映像信号を第2出力として出力する時間軸変換手段を有し、該第1出力からの信号を前記第1チャンネルの映像信号とすると共に、該第2出力からの信号を前記第2チャンネルの映像信号とするようにしたことを特徴とする映像信号の伝送装置。

7. 特許請求の範囲第5項に記載の伝送装置において、互いに視差の異なる2つの立体映像情報に基づく2つの映像信号のうち、一方を前記第1チャンネルの映像信号とし、他方を第2チャンネルの映像信号としたことを特徴とする映像信号の伝送装置。

8. 映像信号を伝送するための伝送装置において、該映像信号から、輝度信号の低域成分と高域成分、及び2つの色度信号のうちの少なくとも一方の色度信号の低域成分と高域成分とを分離し、前記輝度信号の低域成分と色度信号の低域成分とを時分割多重して出力する第1の信号変換手段と、分離された前記輝度信号の高域成分と色度信号の高域成分とを時分割多重し、その時分割多重された信号を低域側に周波数変換して出力する第2の信号変換手段と、該第1及び第2の信号変換手段からの各出力信号を入力し、各々の信号におけるフィールド内の時間的に隣接するライン同士、あるいは隣接するフィールド間あるいはフレーム間の空間的に隣接するライン同士のうち、一方のラインにおいては、前記第1及び第2の信号変換手段からの各出力信号の和に相当する位相関係でその両者を周波数多重し、もう一方のラインにおいては、前記第1及び第2の信号変換手段からの各出力信号の差に相当する位相関係でその両者を周波数多重する周波数多重手段と、から成り、該周波数多重手段によって多重された信号を伝送するようにしたことを特徴とする映像信号の伝送装置。

3. 発明の詳細な説明

（産業上の利用分野）

本発明は、複数のチャンネルあるいは広帯域の映像信号を限られた伝送帯域で伝送するのに好適な映像信号の伝送方法とその装置に関するものである。

尚、ここでいう伝送とは広い意味での伝送であり、例えば、記録・再生も一種の伝送ということで、この伝送という言葉の範ちゅうに含まれる。但し、以下の文中において、場合によっては、伝送と記録・再生とを分けて考える場合もあり、その様な場合、伝送という言葉は送受信機間の伝送などの様な狭い意味で用いられる。

（従来の技術）

近年では、現行のテレビ方式に比して格段の高精細度、高画質の得られるいわゆる高品位テレビのような、新しい高精細テレビ方式の検討が進められており、この高精細テレビ方式では、現行のテレビ方式におけるよりも数倍の画像情報を有し、従って数倍の広帯域を必要とする。また他方では、映像を3次元的に表示させる立体テレビ方式の検討も進められており、この立体テレビ方式では互いに視差の異なる2種類の映像信号を伝送する必要から、伝送路の信号伝送容量として2チャンネル分が必要となる。

以上のように、高精細テレビあるいは立体テレビなどの新しいテレビ方式では、広帯域あるいは複数チャンネルの伝送路が必要となるため、帯域あるいはチャンネル数の制限されている既存の伝送チャンネルで、こうした新しいテレビ方式のサービスを行うためには、広帯域あるいは複数のチャンネルの映像信号を1チャンネル分の限られた伝送帯域で伝送する必要がある。

また、こうした新しいテレビ方式で得られる映像信号を、ビデオ・テープ・レコーダ（VTR）やビデオ・ディスク・プレーヤ（VDP）などで記録し再生する場合を考えて見ても、記録・再生すべき映像信号が広帯域あるいは複数チャンネルの信号であれば、通常の映像信号を記録・再生する場合に比べ記録容量が大きくなってしまうが、1チャンネル分の限られた伝送帯域の信号となれば、記録容量が大きくなることもない。

そこで、この様な1チャンネル分の限られた伝送帯域で信号を伝送する方法として、従来では、例えば、テレビジョン学会技術報告Vol.7, No.44（1984年3月）TEBS95-2における二宮，大塚，和泉による“高品位テレビの衛星1チャンネル伝送方式（MUSE）”と題する文献において論じられているものなどがある。

しかし、この既提案例では独立した2つのチャンネルの映像信号を同時に1つのチャンネルで伝送し、あるいは記録・再生する技術については開示されておらず、従って、こうした装置の実現が重要な課題となっている。

〔発明が解決しようとする問題点〕

上記した様に、従来技術では、広帯域あるいは複数チャンネルの映像信号を1チャンネル分の限られた伝送帯域で伝送させることが完全にはできず、従って、高精細テレビあるいは立体テレビなどの新しいテレビ方式のサービスを行うことが困難であった。また、こうした新しいテレビ方式で得られる映像信号を、VTRやVDPなどで記録・再生する場合においても、記録・再生すべき映像信号が広帯域あるいは複数チャンネルの信号のままであれば、記録容量が増大してしまって、記録媒体の記録容量が限られている場合には、長時間に亘る録画再生が行えないという問題があった。

本発明は、上記した従来技術の問題点に鑑みなされたものであり、従って、本発明の目的は、広帯域あるいは複数のチャンネルの映像信号を1チャンネル分の帯域で伝送あるいは記録・再生できる映像信号の伝送方法およびその装置を提供することにある。

〔問題点を解決するための手段〕

本発明は、上記目的を達成するために、伝送すべき第1チャンネルの映像信号V_1と第2チャンネルの映像信号V_2とを、第1の信号ブロック（例えば偶数番目のラインで構成されるブロック）では、上記第1チャンネル映像信号V_1と第2チャンネル映像信号V_2との和（V_1+V_2）に相当する位相関係をもって互いに周波数多重して伝送し、上記第1の信号ブロック以外の第2の信号ブロック（例えば奇数番目のラインで構成されるブロック）では、上記第1チャンネル映像信号V_1と第2チャンネル映像信号V_2との差（V_1-V_2）に相当する位相関係をもって互いに周波数多重して伝送するようにしたものである。

〔作用〕

上記により、第1チャンネル映像信号V_1と第2チャンネル映像信号V_2は、1チャンネル分の帯域内で互いに周波数多重されるため、2チャンネルの映像信号（V_1とV_2）を1チャンネル分の帯域で伝送することができる。

また、上記のようにして周波数多重された映像

信号のうち、上記第1の信号ブロックに相当する映像信号(V_1+V_2)と、上記第2の信号ブロックに相当する映像信号(V_1-V_2)と、で和分演算をすれば、上記第1チャンネルの映像信号V_1が分離抽出され、また、その両者で差分演算をすれば、上記第2チャンネル映像信号V_2が分離抽出され、かくして2チャンネルの映像信号V_1とV_2が復元される。

〔実施例〕

以下、本発明の実施例を図面により説明する。

第1図は、2つのチャンネルの映像信号を1つのチャンネルの映像信号に変換して伝送する、本発明の一実施例としての伝送装置を示すブロック図であり、第2図は上記映像信号を水平走査線単位で表示した説明図である。

第1図において、1は第1のチャンネルの映像信号V_1が入力される端子、2は第2のチャンネルの映像信号V_2が入力される端子、3はこれら第1及び第2のチャンネルの映像信号V_1とV_2が1つのチャンネルに合成されて出力される端子である。また、10は位相反転回路、20は切換回路、30は合成回路である。

第1チャンネルの映像信号V_1と第2チャンネルの映像信号V_2は、互いに同期した位相関係でそれぞれ端子1と2に供給される。一例として、第2図に示すように映像信号の水平走査線単位でラインL_1, L_2, L_3, …の順で上記第1チャンネルの映像信号V_1が端子1に入力されるのに対して、それと同期してラインL_1, L_2, L_3, …の同じ順序で上記第2チャンネルの映像信号V_2が端子2に入力される。端子2からの映像信号V_2は切換回路20の一方の端子A側に供給されると共に、位相反転回路10にも供給され、そして、この位相反転回路10にて位相反転された出力が、切換回路20の他方の端子B側に供給される。

この切換回路20にて、入力映像信号V_2(あるいはV_2)の水平走査線単位で端子A側とB側とが交互に切換えられ、その出力は合成回路30に供給され、そして、端子1から供給された上記第1チャンネルの映像信号V_1と上記切換回路20からの出力映像信号とがこの合成回路30にて加算されて合成される。ここで、合成回路30は、少なくとも映像表示期間(つまり、同期信号部分を含まない期間)では単なる加算器であると考えて良い。

従って、この合成回路30からは、第2図に示すように、ラインL_1の期間では、第1チャンネルの映像信号V_1と第2チャンネルの映像信号V_2の和成分(V_1+V_2)$_1$が出力され、次のラインL_2の期間では、第1チャンネルの映像信号V_1と第2チャンネルの映像信号V_2との差成分(V_1-V_2)$_2$が出力される。一般には、2n番目のラインでは、第1チャンネルの映像信号V_1と第2チャンネルの映像信号V_2との和成分(V_1+V_2)$_{2n}$が出力され、次の(2n+1)番目のラインでは、第1チャンネルの映像信号V_1と第2チャンネルの映像信号V_2との差成分(V_1-V_2)$_{2n+1}$が出力される。即ち、以上の様な合成回路30における加算演算によって、第1チャンネルの映像信号V_1と第2チャンネルの映像信号V_2とが和または差に相当する位相関係で周波数多重されるわけである。

以上により、第1及び第2の2つのチャンネルの映像信号は、1つのチャンネルの映像信号V_0に変換されて、端子3より出力される。

出力映像信号V_0は、以上の説明から明らかなように、入力映像信号V_1とV_2との和成分または差成分であるので、この出力映像信号V_0の占有帯域は入力映像信号V_1あるいはV_2のいずれか帯域の広い方で決まり、上記第1及び第2チャンネルの映像信号の占有帯域が同じでその値をBWとすれば、上記出力映像信号V_0の占有帯域も同じBWとなる。これを換言すれば、2チャンネル分の映像信号を伝送するのに必要な帯域(2×BW)に対し、本実施例によれば、その半分の帯域(BW)で2チャンネル分の映像信号を伝送できることになる。

なお、上記2n番目のライン(第2図の実線で示すラインL_{2n})と上記(2n+1)番目のライン(第2図の破線で示すラインL_{2n+1})との位相

関係は、フィールド内の時間的に隣接するライン同士をさす場合の他、例えば実線のラインL_{2n}を第1フィールド（あるいは第1フレーム）内のラインとし、破線のラインL_{2n+1}をこの第1フィールド（あるいは第1フレーム）に続く次の第2フィールド（あるいは第2フレーム）内のラインとするような、フィールド間あるいはフレーム間の空間的に隣接するライン同士をさす場合であっても良く、いずれの場合も本発明の範ちゅうに含まれる。

次に、上記の如くして1つのチャンネルに合成された映像信号V_3より元の映像信号V_1とV_2を分離抽出するための本発明に係わる信号逆変換装置の一具体例を第3図に示す。

同図において、上記第1図の実施例に示す伝送装置により出力されて所定伝送路を伝送され、あるいは前記VTRやVDPのように記録信号処理されて記録・再生されて得られる上記映像信号V_3は、入力端子4に供給される。ここで、50は加算器、60は減算器であり、また、40は、端子4からの入力映像信号を、上記のフィールド内あるいはフィールド間あるいはフレーム間で時間的あるいは空間的に隣接する2つのラインL_{2n}とL_{2n+1}との位相差に相当する時間T（第2図のT）だけ遅延する遅延回路である。

上記加算器50にて、入力端子4からの入力映像信号V_3と、それを遅延回路40にて時間Tだけ遅延して出力される映像信号V_3'とが加算される。従って、一般にライン番号2nの期間では、端子4からのライン番号2nの映像信号$(V_1+V_2)_{2n}$と、それより1つ前のライン番号（2n−1）の映像信号$(V_1-V_2)_{2n-1}$とが、この加算回路50で加算されて、次式で表現される映像信号$(V_1')_{2n}$がこの加算回路50より出力される。

$$(V_1')_{2n} = (V_1+V_2)_{2n} + (V_1-V_2)_{2n-1} \quad \cdots\cdots \quad (1)$$

同様に、次のライン番号（2n+1）の期間では、次式で表現される映像信号$(V_1')_{2n+1}$がこの加算回路50より出力される。

$$(V_1')_{2n+1} = (V_1-V_2)_{2n+1} + (V_1+V_2)_{2n} \quad \cdots\cdots \quad (2)$$

次に、上記減算器60にて、端子4からの入力映像信号V_3と、それを遅延回路40にて時間Tだけ遅延して出力される映像信号V_3'とが減算され、従って一般に、ライン番号2nの期間では、端子4からのライン番号2nの映像信号$(V_1+V_2)_{2n}$と、それより1つ前のライン番号（2n−1）の映像信号$(V_1-V_2)_{2n-1}$とが、この減算回路60で減算されて、次式で表現される映像信号$(V_2')_{2n}$がこの減算回路60より出力される。

$$(V_2')_{2n} = (V_1+V_2)_{2n} - (V_1-V_2)_{2n-1} \quad \cdots\cdots \quad (3)$$

同様に、次のライン番号（2n+1）の期間では、次式で表現される映像信号$(V_2')_{2n+1}$がこの減算回路60より出力される。

$$(V_2')_{2n+1} = (V_1-V_2)_{2n+1} - (V_1+V_2)_{2n} \quad \cdots\cdots \quad (4)$$

一般に映像信号は、フィールド内あるいはフィールド間あるいはフレーム間で時間的あるいは空間的に隣接する1つのライン間（L_{2n}とL_{2n-1}の間、あるいはL_{2n+1}とL_{2n}の間）では、強い相関を有するため、ライン間で逆相の関係にある成分、即ち上記(1)式及び(2)式のV_2の成分、及び上記(3)式及び(4)式のV_1成分は、互いに打ち消し合ってほぼ零とみなすことができる。従って、上記(1)式及び(2)式から明らかなように、上記加算回路50からは、第2チャンネルの映像信号V_2の成分は除去され、第1チャンネルの映像信号V_1成分のみが分離抽出され端子5より出力される。同様に、上記(3)式及び(4)式から明らかなように、上記減算回路60からは第1チャンネルの映像信号V_1の成分は除去され、第2チャンネルの映像信号V_2成分のみが出力される。

この第3図の逆変換装置では、上記(1)式及び(2)式から明らかなように、加算回路50から分離出力される映像信号V_1は、ライン番号2nと次のライン番号（2n+1）とで、共に同相となる。これに対し、上記(3)式及び(4)式から明らかなように、減算回路60から分離出力される映像信号V_2

は、ライン番号2nと次のライン番号（2n＋1）とで、互いに逆相となる。

そこで、上記減算回路60からの出力映像信号は切換回路80の端子A側に供給されると共に、位相反転回路70に供給され、そして、この位相反転回路70にて位相反転された出力は、切換回路80の他方の端子B側に供給される。この切換回路80は周期T毎に端子A側とB側に交互に切換えられる。即ち、具体的には、上記ライン番号2nの期間では、端子A側に接続され、次のライン番号（2n＋1）の期間では、端子B側に接続され、以上の切換が上記周期T毎に交互に行われる。その結果、上記切換回路80からは、ライン番号2nと次のライン番号（2n＋1）とで共に同相となる映像信号が出力され端子6に出力される。

かくして、2チャンネルの映像信号が1つのチャンネルに合成された入力映像信号V_0より、第1チャンネルの映像信号V_1'と第2チャンネルの映像信号V_2'とが分離され、かつ元の位相関係も正しく復元されて、それぞれ端子5と6より出力される。

ところで、第1図に示した実施例では上記したように、画像の有する相関を利用して2つのチャンネルの映像信号を多重するものであり、相関のない信号を多重すると相互に妨害を与える問題を生ずる。そこで、次の実施例では、こうした問題を生じ難いようにするために、多重すべき信号に制約を与えるものである。

第4図は本発明の他の実施例を示すブロック図、第5図は第4図における各部信号のタイミングチャート、である。

一般に映像情報の伝送には、明暗を表わす輝度情報と、色彩を表わす色度情報が必要であり、輝度情報と色度情報の間には相関はない。そこで、本実施例では、2つのチャンネルの映像信号を多重するに際し、各映像信号を輝度信号と色度信号とに分離して時分割で多重し、第1チャンネル映像信号の輝度信号には第2チャンネル映像信号の輝度信号を周波数多重し、第1チャンネル映像信号の色度信号には第2チャンネル映像信号の色度信号を周波数多重するようにしたものである。この方法により、互いに相関のない輝度信号と色度信号が時間的に重ならないようにでき、従って相互の妨害をなくすことができる。

では、第4図及び第5図を用いて、本実施例を更に詳しく説明する。

本実施例は、先の第1図の実施例における入力端子1と2に時分割信号処理回路100を挿入して構成される。尚、第4図において、第1図と同じ回路ブロックには同じ符号を付してあり、その動作は第1図とまったく同じであるので、説明は省略する。

第4図において、端子1'には第1チャンネルの映像信号E_1が、端子2'には第2チャンネルの映像信号E_2がそれぞれ入力される。

時分割信号処理回路100において、端子1'からの第1チャンネル映像信号E_1は、その水平走査線単位で第5図のa～cに示すように、輝度信号Y_1と2つの色度信号C_{R1}とC_{B1}とに分離され、その後、各々時間軸圧縮されて時分割で多重されて、第5図のcに示す様な映像信号V_1として端子1より出力され、上記合成回路30に供給される。

同様に、端子2'からの第2チャンネル映像信号E_2は、その水平走査線単位で第5図のd～fに示すように、輝度信号Y_2と2つの色度信号C_{R2}とC_{B2}とに分離され、その後、各々時間軸圧縮されて時分割で多重されて、第5図のhに示す様な映像信号V_2として端子2より出力され、上記切換回路20と位相反転回路10とに供給される。

上記映像信号V_1とV_2は、それぞれの輝度信号（Y_1とY_2）と2つの色度信号（C_{R1}とC_{R2}、及びC_{B1}とC_{B2}）が同じタイミング関係で出力される。上記回路10，20，30にて、上記第1図で述べたと同じ信号処理が行われ、1つのチャンネルの映像信号V_3として端子3より出力される。

従って、第5図のiに示すように、一般にライン番号2nの期間では、第1チャンネルの映像信

号V_1と第2チャンネルの映像信号V_2との輝度信号同士の和成分（Y_1+Y_2）、及び色度信号同士の和成分（$C_{11}+C_{21}$）と（$C_{12}+C_{22}$）とが時分割多重された形態で映像信号（V_1+V_2）$_{2n}$として端子3より出力される。同様に、次のライン番号（2n+1）の期間では、第5図のbに示すように、第1チャンネルの映像信号V_1と第2チャンネルの映像信号V_2との輝度信号同士の差成分（Y_1-Y_2）、及び色度信号同士の差成分（$C_{11}-C_{21}$）と（$C_{12}-C_{22}$）とが時分割多重された形態で映像信号（V_1-V_2）$_{2n+1}$として端子3より出力される。

次に、以上の様にして1つのチャンネルに合成された映像信号V_0より、上記第1及び第2チャンネルの映像信号V_1とV_2を分離するためには、前述の第3図に示した信号逆変換装置が同様に適用できる。この第3図の信号逆変換装置を適用した場合、端子5及び6より出力される映像信号V_1'及びV_2'は、第5図のa及びbに示す波形とほぼ同様の、輝度信号と色度信号が時分割多重された形態の信号となる。

従って、これより元の映像信号E_1とE_2を復元するための信号処理回路が、図示しないが、上記第3図の出力端子5，6に更に接続される。即ち、この信号処理回路において、上記映像信号V_1'とV_2'のそれぞれより、時分割多重された輝度信号と色度信号がそれぞれ分離され、かつ元の正規の時間軸を有するようにそれぞれ適宜時間軸処理されて、その結果、元の映像信号E_1とE_2に準ずる信号がそれぞれ出力される。

尚、ここで、元の映像信号E_1とE_2と全く同様の信号が得られるのではなく、それらに準じた信号が得られるのは、この信号処理回路に入力される映像信号V_1'とV_2'が第5図a及びbに示した映像信号V_1，V_2と完全には一致していないからである。

次に、伝送すべき映像信号が、映像を3次元的に表示させるための立体映像信号である場合と、映像の細部を表示させるための高精細映像信号である場合とについてそれぞれ説明する。

先ずは、立体映像信号についてである。立体映像情報として、一般には、右眼用と左眼用の互いに視点の異なる2つの映像情報が必要である。従って、この立体映像信号を本発明を用いて伝送する場合は、互いに視点の異なる映像情報に基づく第1及び第2の立体映像信号をそれぞれ上記第1及び第2のチャンネルの映像信号として、上記第1図あるいは上記第4図の伝送装置の入力端子に供給するようにすれば良い。上記2つの立体映像信号は、その性質から、一般的には、信号間に強い相関があるため、チャンネル間の信号漏洩の影響は大幅に低減され、効率良く伝送することができる。

次は、高精細映像信号についてである。高精細映像信号として、一般には広帯域の映像信号が必要である。従って、この高精細映像信号を本発明を用いて伝送する場合は、第6図に示す様な伝送装置を用いれば良い。

第6図は本発明の別の実施例を示すブロック図、第7図は第6図における各部信号のタイミングチャートである。

第6図において、200は時間軸変換回路であり、他のブロックは上記第1図と同じであり、同一符号を付してある。

端子3'に入力される高精細映像信号E_0は、時間軸変換回路200にて、水平走査線単位で適宜時間軸伸長され、例えば2倍に伸長されて、2倍の周期T_W（T_W＝2H：Hは1水平走査周期）で、第7図のaに示すように一般にライン番号4nでは、輝度信号Y_1と色度信号C_1とが時分割多重されて、映像信号V_1として端子1より出力され、次のライン番号（4n+1）では第7図のbに示すように、輝度信号Y_2と色度信号C_2とが時分割多重されて、信号V_2として端子2より出力される。上記映像信号V_1とV_2は、それぞれの輝度信号（Y_1とY_2）と色度信号（C_1とC_2）が同じタイミングで出力される。その結果、端子3からは、第7図のcに示すように上記ライン4n番目の映像信号V_1とライン（4n+1）番目の映像信号V_2との輝度信号同士の和成分（

Y_1+Y_2）、及び色度信号同士の和成分（C_1+C_2）とが時分割多重された形態でライン番号2nの映像信号（V_1+V_2)$_{2n}$として出力される。

同様に、次のライン番号（4n+2）では、第7図のcに示すように輝度信号Y_1と色度信号C_1とが時分割多重で映像信号V_1として端子1より出力され、次のライン番号（4n+3）では、第7図のeに示すように輝度信号Y_2と色度信号C_2とが時分割多重で映像信号V_2として端子2より出力される。従って、端子3からは、第7図のfに示すように上記ライン（4n+2）番目の映像信号V_1とライン（4n+3）番目の映像信号V_2との輝度信号同士の差成分（Y_1-Y_2）、及び色度信号同士の差成分（C_1-C_2）とが時分割多重された形態でライン番号（2n+1）の映像信号（V_1-V_2)$_{2n+1}$として出力される。

次に、以上の様にして、1つのチャンネルに合成された映像信号V_3より、映像信号V_1とV_2を分離するためには、前述の第3図に示した信号逆変換装置が適用できる。即ち、映像信号V_3は、上記第3図の信号逆変換装置の入力端子4に供給され、端子5及び6からは上記第7図のc及びe（あるいはd及びe）に示す波形とほぼ同様の形態の、輝度信号と色度信号が時分割多重されたライン間隔T_0の映像信号V_1'とV_2'が出力される。

従って、これより元の映像信号E_1を復元するための時間軸逆変換回路が、図示しないが、上記第3図の信号逆変換装置の出力端子5，6に更に接続される。即ち、この時間軸逆変換回路において、上記映像信号V_1'とV_2'のそれぞれより時分割多重された輝度信号と色度信号がそれぞれ分離され、かつ元の正規の時間軸を有するようにそれぞれ適宜時間軸処理されて、その結果、元の高精細映像信号E_1に準ずる信号が出力される。

以上第6図の実施例によれば、広帯域を必要とする高精細映像信号E_1を、時間軸伸長によって、例えば2倍の時間伸長により1／2の占有帯域でかつ1チャンネルで伝送できる効果が得られる。

次に、上記高精細映像信号に適用する本発明の更に別の実施例を第8図に示す。同図において、300は信号変換回路であり、他のブロックは上記第1図と同じであり、同一符号を付してある。また第9図は、第8図における各部信号のタイミングチャートである。

端子3′に入力される高精細映像信号E_0は、信号処理回路310にて、水平走査線単位で、輝度信号Yと色度信号Cとに分離されて時分割で多重され、かつその時分割多重された信号は第9図のaとbに示す様に低域信号成分と高域信号成分との2つに分割される。

一方の低域信号成分は、第9図のaに示すように低域映像信号V_1として端子1より出力される。この低域映像信号V_1は、上記より明らかなように、色度信号Cの低域成分C_Lと輝度信号Yの低域成分Y_Lとが時分割多重された形態を有する。

上記他方の高域信号成分は、上記低域映像信号V_1と占有帯域がほぼ等しくなるように、あるいは占有帯域幅が上記低域映像信号V_1のそれより小さくなるように、周波数変換回路320にて周波数変換され、第9図のbに示すように高域映像信号V_2として端子2より出力される。この高域映像信号V_2は、上記より明らかなように、色度信号Cの高域成分C_Hと輝度信号Yの高域成分Y_Hとが時分割多重されて周波数変換された形態を有する。

上記低域映像信号V_1と高域映像信号V_2は、それぞれの輝度信号（Y_LとY_H）と色度信号（C_LとC_H）が、同じタイミングで出力される。

以上により端子3からは、第9図のcに示すように、一般にライン番号2nの期間では、上記低域映像信号V_1と高域映像信号V_2との輝度信号同士の和成分（Y_L+Y_H）、及び色度信号同士の和成分（C_L+C_H）とが時分割多重された形態で映像信号（V_1+V_2)$_{2n}$として出力される。同様に、次のライン番号（2n+1）の期間では、第9図のdに示すように、上記低域映像信号V_1と高域映像信号V_2との輝度信号同士の差成分（Y_L-Y_H）、及び色度信号同士の差成分（C_L-C_H）とが時分割多重された形態で映像信号（V_1-V_2)$_{2n+1}$として端子3より出力される。

次に、以上の様にして、1つのチャンネルに合成された映像信号V_cより映像信号V_aとV_bを分離するためには、前述の第3図に示した信号逆変換装置が適用できる。即ち、映像信号V_cは、上記第3図の信号逆変換装置の入力端子4に供給され、端子5及び6からは、上記第9図のa及びbに示す波形とほぼ同様の波形の、輝度信号と色度信号が時分割多重された映像信号V_1'とV_2'がそれぞれ出力される。

従って、これより元の映像信号E_0を復元するための信号逆変換回路500が上記第3図の信号逆変換装置の出力端子5，6に更に接続される。

この信号逆変換回路500の一具体例を第10図に示す。同図において、端子6'に入力される上記第3図の端子6からの高域映像信号V_2'は、周波数変換回路510にて元の占有帯域を有するように周波数変換される。そして、信号処理回路520にて、上記周波数変換回路510からの出力と、端子5'に入力される上記第3図の端子5からの低域映像信号V_1'とが合成され、その合成結果より上記の時分割多重された輝度信号Yと色度信号Cがそれぞれ分離され、そして、元の正規の時間軸を有するようにそれぞれ時間軸処理されて、元の高精細映像信号E_0に準ずる映像信号E_0'が端子7より出力される。

なお、上記第8図あるいは第6図における入力映像信号E_0及び上記第10図における出力映像信号E_0'の信号形態として、上記輝度信号Yと色度信号Cが周波数多重された、いわゆるコンポジット信号の形態でも良いが、それ以外に上記輝度信号Yと色度信号Cをそれぞれ個別に入出力させる形態でも良く、あるいは赤，緑，青のいわゆる三原色映像信号をそれぞれ個別に入出力させるような形態でも良く、いずれも本発明の範ちゅうに含まれるものである。

また、上記第6図，第8図，第10図では、色度信号として上記Cを用いて説明しているが、上記第4図の実施例で示したように、一般には色度情報として2つの色度信号（C_wとC_n）が必要であり、上記色度信号Cはこの2つの色度信号を表わすものである。

また、本発明は、この2つの色度信号を伝送する方法として、上記第4図で述べたような水平走査線単位で常に2つの色度信号を伝送する、いわゆる同時式の場合に適用できるが、本発明はこれに限定されるものではなく、上記2つの色度信号を水平走査線単位で交互に伝送する、いわゆる線順次式の場合にも適用できるものである。

特にこの線順次式で伝送する場合においては、上記第6図及び第8図の実施例における2つのチャンネル間の色度信号の和（C_1+C_2、あるいはC_3+C_4）及び差（C_1-C_2、あるいはC_3-C_4）の演算はいずれも上記2つの色度信号（C_wとC_n）のいずれか一方同士の演算、即ち、C_w同士の和・差演算かあるいはC_n同士の和・差演算で行われる。一般には、上記2つの色度信号（C_wとC_n）の間には相関はないが、同じ色度信号同士（C_w同士あるいはC_n同士）には強い相関があるため、上記の本発明の方法によれば、線順次式の場合においても2つのチャンネル間における色度信号相互の妨害を大幅に改善できることが容易に理解されるであろう。

次に、上記第6図及び第10図における周波数変換回路110及び510における周波数変換は、これらに入力される映像信号と図示しないが内部に有する局部発振器からの局発信号との乗算を行い、両者の差周波数成分を抽出することにより達成される。

ここで、上記第8図及び第10図の回路構成は、この局発信号の位相が隣接ライン間（ライン番号2nと2n+1の間）で同相となるように設定した場合を示したものである。従って、この局発信号の位相が隣接ライン間で逆相となるように設定した場合は、上記第8図の位相反転回路10と切換回路20は不要となり、上記周波数変換回路110の出力を直接、上記合成回路30に供給すれば良く、また上記第10図の位相反転回路70と切換回路80も不要となり、上記減算回路50からの出力を直接、上記第10図の周波数変換回路510に供給すれば良く、この場合においても得ら

れる効果に同じて、本発明の主旨にそうものである。

また、上記第5図，第7図，第9図に示す例においては、同期信号部を示し、この同期信号に対しても上記映像信号と同様の和・差演算を施しても良いが、それを施さなくても良く、本発明の主旨をそれるものではない。即ち、同期信号に対して上記和・差演算を施せば、和の施された（番号2nの）ラインと、差の施された（番号2n+1の）ラインとで波形の異なる同期信号が得られるため（例えば、和演算では、2倍の振幅を有する同期信号が得られ、差演算では、振幅0の同期信号が得られる。）、その波形の違いを検出することにより和の施されたラインであるか、差の施されたラインであるかを検知できる副次的効果が得られる。また逆に、同期信号に対して上記和・差演算を施さなければ、各ラインで一様の波形を有する同期信号を得ることができ、全てのラインで同期信号を安定に検出できる副次的効果が得られる。

（発明の効果）

以上述べたように、本発明によれば、複数のチャンネルあるいは広帯域の映像信号を限られた伝送帯域でチャンネル間の信号混信の影響を大幅に低減して、効率良く伝送あるいは記録・再生することができる。従って、既存の伝送路を用いて、広帯域あるいは複数チャンネルを必要とする高品位テレビあるいは立体テレビ等の新しいテレビ方式のサービスを行うことができ、またこれらの新しいテレビ方式に対応するVTRやVDPのような映像信号記録再生装置においては、実質的に高密度記録を実現することができ（即ち、従来の1チャンネル分の映像信号を記録するための記録容量と同じ容量で、複数チャンネルあるいは広帯域の映像信号の情報を記録できるからである。）、録画再生時間の長時間化を容易に達成できる効果が得られる。

4. 図面の簡単な説明

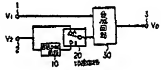
第1図は本発明の一実施例を示すブロック図、第2図は映像信号を水平走査線単位で示した説明図、第3図は本発明に係わる信号変換装置の一具体例を示したブロック図、第4図は本発明の他の実施例を示すブロック図、第5図は第4図における各部信号のタイミングチャート、第6図は本発明の別の実施例を示すブロック図、第7図は第6図における各部信号のタイミングチャート、第8図は本発明の更に別の実施例を示すブロック図、第9図は第8図における各部信号のタイミングチャート、第10図は第3図の信号逆変換装置に使用される信号逆変換回路の一具体例を示すブロック図、である。

符号の説明

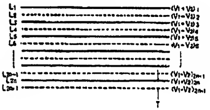
10，70…位相反転回路、20，80…切換回路、30…合成回路、40…遅延回路、50…加算器、60…減算器、100…時分割信号処理回路、200…時間軸変換回路、300…信号変換回路、500…信号逆変換回路

代理人 弁理士 並木 昭夫

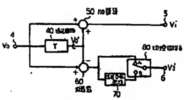
第1図



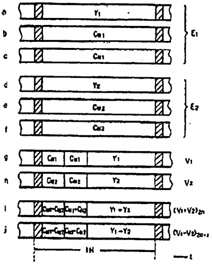
第2図



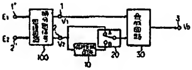
第3図



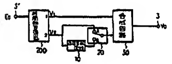
第5図



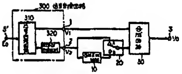
第4図



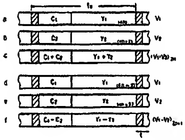
第6図



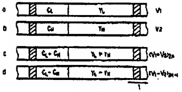
第8図



第7図



第9図



第10図